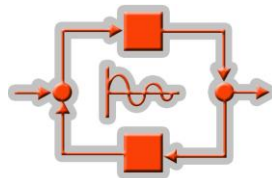




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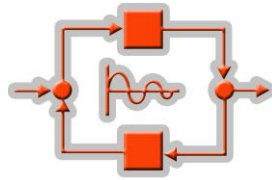
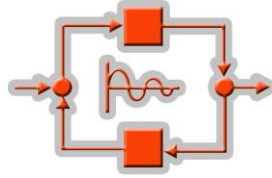


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Controller Design for Magnetic Levitation System Using Diophantine Minimum Order Estimator Approach

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Abstract

This research paper presents controller design for Magnetic Levitation System using modern discrete time control theory approach. Zero-Order-Hold (ZOH) method is used to discretize the Jacobian linearization state space magnetic levitation dynamic model. This model is used to design minimum order state estimator using Diophantine's Polynomial approach which implicitly implements state estimation. Pole assignment is used for state feedback. Response of the controller and estimator is kept Dead-Beat. Difference equation of the designed controller is implemented on digital controller interfaced with digital computer for real time monitoring. Simulation and hardware testing results show that this approach is intuitive, simpler, gives good transient control performance and superior overall response of the system to the noise.¹

Keywords: Diophantine state estimator, minimum order polynomial observer, Dead beat, Pole assignment, Magnetic levitation

Nomenclature

MagLev	Magnetic Levitation
EM	Electromagnet
[M]	Diophantine Estimator Coefficient Matrix
[E]	Sylvester's Matrix
NI™	National Instruments
DAQ	Data Acquisition Card
ADC	Analog to Digital Converter
PWM	Pulse Width Modulation
SISO	Single Input Single Output
MOS	Metal Oxide Semiconductor
FET	Field effect transistor

1. Introduction

Magnetic Levitation is a challenging system and inherently nonlinear and unstable without feedback [1, 2]. It involves suspending a ferromagnetic object such as a steel ball in space. This setup acts as a benchmark for

research of Maglev trains that move more smoothly and somewhat more quietly than wheeled mass transit systems. The power needed for levitation is typically not a large percentage of the overall energy consumption [3]. Levitation stability can be achieved but most of the controllers give poor reference position tracking response of the levitated object which is one of the main issues addressed in this paper along with employing a different technique of polynomial control for digital controller design. Anticipatory control techniques give poor response to system noises.

Actuation in MagLev systems is mostly achieved by an electromagnet. Current in the field coil is adjusted to generate a variable magnetic field strength in response to the position of the ball. Maglev is an area of active research due to its significance for future transport technology and applications, such as contact-less magnetic bearings and conveyor mechanisms [4, 5].

This paper implements polynomial equation approach for controller design of Maglev system. This technique is alternative synthesis approach of controller design to state space and comprises of polynomial algebra rather than working with system matrices. This technique can also be used to synthesize and implement other estimators such as Kalman filter, model matching, and linear Gaussian quadratic (LGQ) etc. Polynomial approach gives results directly in the form of difference equations and implements state observation implicitly rather than explicitly as in state space approach, hence this technique is quicker and simpler [6, 7]. Minimum order polynomial observer estimates only unmeasured states in model. Diophantine equations are used to synthesize polynomial estimator. Dead-beating is implemented for fastest possible response. Simulation results are presented. A hardware platform is designed and tested.

Results of controller based on polynomial approach show a very good system behavior for tracking square, sinusoidal and trapezoidal reference inputs of reasonable magnitudes and frequencies up to 15Hz. Current technique is not optimal for any performance index, our main objective is reference tracking and fastest response

¹ This research work has been conducted in Control Systems laboratory at COMSATS IIT, Wah Campus.



for which dead-beat pole assignment is done. However response is tested against added white Gaussian noise of 500Hz bandwidth to reference inputs and response is found to be stable.

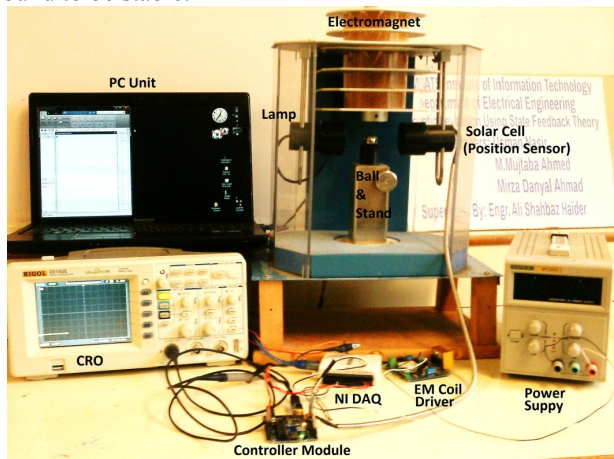


Figure 1. Hardware Setup

This paper has been organised as follows, section 2 describes the hardware setup design. Dynamic modelling of the system is described in Section 3. Controller is designed in section 4. Simulations and hardware results are presented in section 5.

2. Electromechanical Design

Mechanical design: Hardware platform of the system is shown in figure. 1. It consists of an electromagnets and a steel ball. In free air ball drops down under the action of gravity so electromagnet is used to provide counter gravity force. Electromagnet consists of 0.75mm diameter copper wire wound by 3500 turns. Cross section of the core is a square of dimension 20x20mm made up of steel laminations. Steel balls weight may vary from 30g to 100g. A bulb and solar panel assembly is used to sense the position of the steel ball in air.

Operating principle of the system: System operates as follows. Light bulb and solar panel are placed opposite to each other in parallel as shown in figure. 2. The amount of light of the bulb incident on the solar panel is a function of position of the ball between them. If the ball moves up the shaded area of the panel increases and its output decreases and vice versa. This sensor output after signal conditioning is fed to microcontroller and is used in control law to generate the actuation signal to drive the electromagnet in response to the position of the ball in air.

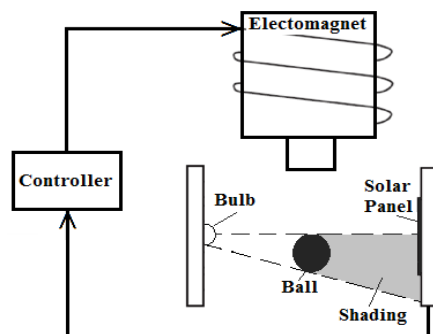


Figure 2. Operating Principle.

Electrical design: Electromagnet is operated at 32VDC with a maximum current of 1.5A. This gives electromagnet enough power to lift a ball of mass 100g from a distance of 55mm. Resistance of the electromagnet coil is 22Ω and inductance is 0.16H without presence of the steel ball.

3. System Dynamics

Non-linear differential equation model system: Free body diagram of the system is shown in figure. 3. Steel ball is acted upon by two forces namely downwards force of gravity and upward magnetic pull $f(t)$.

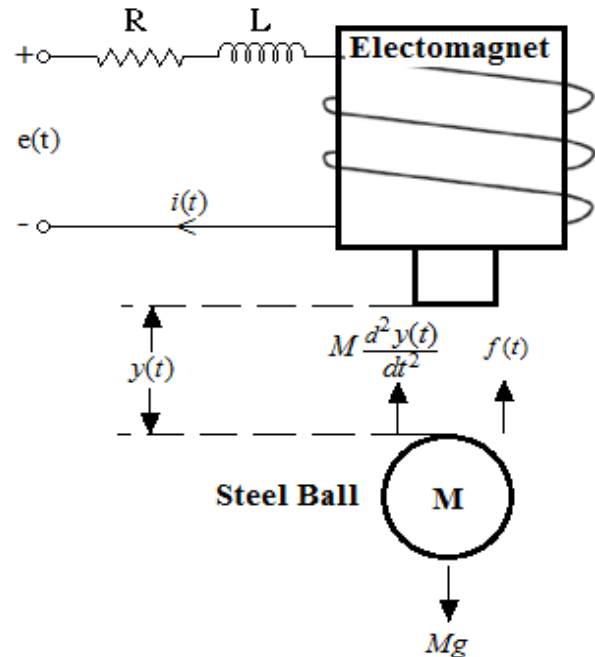


Figure 3. Free body diagram and electrical circuit

If we neglect the air friction then according to Newton's second law a second order differential equation describes the mechanical dynamics of the system [1]. This equation is given by,

$$M \frac{d^2 y}{dt^2} = Mg - k \frac{i^2(t)}{y(t)} \quad (1)$$

Here k is a constant. The electrical part of the plant in figure. 3 consists of a series RL circuit which is governed by equation (2),

$$e(t) = L \frac{di(t)}{dt} + i(t) R \quad (2)$$

Equations (1) and (2) give us non-linear differential equation model of the system.

Non-linear state space model: Let's assign the state variables as in equation (3). Here $e(t)$ and $y(t)$ have been chosen as input and output respectively. Hence system is Single Input Single output (SISO).

$$x_1 = y(t), \quad x_2 = \frac{dy}{dt}, \quad x_3 = i(t), \quad u(t) = e(t) \quad (3)$$

Using equations (1), (2) and (3) we get the following State Space equation model,



$$\begin{aligned}
\dot{x}_1 &= x_2 = p_1(t) \\
\dot{x}_2 &= g - \frac{k}{M} \frac{x_3^2}{x_1} = p_2(t) \\
\dot{x}_3 &= -\frac{R}{L} x_3 + \frac{1}{L} u(t) = p_3(t) \\
y &= x_1
\end{aligned} \quad (4)$$

System (4) can be expressed in terms of a non-linear vector function p as follows:

$$\dot{x}(t) = p(x(t), u(t)) \quad (5)$$

Where,

$$p(x(t), u(t)) = \begin{bmatrix} x_2 \\ g - \frac{k}{M} \frac{x_3^2}{x_1} \\ -\frac{R}{L} x_3 + \frac{1}{L} u(t) \end{bmatrix} \quad (6)$$

Linear state space model: System (5) can be linearized around nominal operating point and expressed in terms of deviation variables as given by equation (7),

$$\begin{aligned}
\Delta \dot{x}(t) &= A \Delta x(t) + B \Delta u(t) \\
y &= C \Delta x(t) + D \Delta u(t)
\end{aligned} \quad (7)$$

Nominal operating point is given by,

$$x_{01} = y_0, \quad x_{02} = 0, \quad x_{03} = i_0 \quad (8)$$

From equation (1) we get,

$$k = Mg \frac{x_{01}}{x_{03}^2} = Mg \frac{y_0}{i_0^2} \quad (9)$$

Using equation (2) steady state control input is given by,

$$u_0 = e_0 = i_0 R = \left(\sqrt{\frac{Mgy_0}{k}} \right) R \quad (10)$$

Jacobian linearized model is given by [1],

$$\begin{aligned}
A &= \begin{bmatrix} 0 & 1 & 0 \\ g/y_0 & 0 & -2g/i_0 \\ 0 & 0 & -R/L \end{bmatrix} \\
B &= \begin{bmatrix} 0 \\ 0 \\ 1/L \end{bmatrix}, \\
C &= [1 \quad 0 \quad 0] \\
D &= [0]
\end{aligned} \quad (11)$$

Values of various parameters in (11) are enlisted in Table 1.

Table 1. Parameter values for system in (13).

Parameter	Value
R	22Ω
L	0.16H
g	9.8m/sec ²
y ₀	0.035m
i ₀	1.35A
M	0.06Kg

Substituting the values from Table 1. in equation (11) we get,

$$\begin{aligned}
A &= \begin{bmatrix} 0 & 1 & 0 \\ 280 & 0 & -14.52 \\ 0 & 0 & -137.5 \end{bmatrix} \\
B &= \begin{bmatrix} 0 \\ 0 \\ 6.25 \end{bmatrix} \\
C &= [1 \quad 0 \quad 0] \\
D &= [0]
\end{aligned} \quad (12)$$

Transfer function $C(sI - A)^{-1}B + D$ comes out to be,

$$G_p(s) = \frac{-90.75}{s^3 + 138s^2 - 280s - 38500} \quad (13)$$

Discrete state space model: Zero-Order-Hold method is used to discretize the system in equation (13). Sampling time T_s is chosen to be 2msec. it is kept small owing to the unstable nature of the system. Discrete state space system matrices are given in equation (14).

$$\begin{aligned}
G &= \begin{bmatrix} 1.001 & 0.002 & -2.65e-5 \\ 0.5601 & 1.001 & -0.02539 \\ 0 & 0 & 0.7596 \end{bmatrix} \\
H &= \begin{bmatrix} -1.3e-7 \\ -0.00016 \\ 0.01093 \end{bmatrix}, \quad C = [1 \quad 0 \quad 0] \\
D &= [0]
\end{aligned} \quad (14)$$

System equation (14) is minimal so arbitrary pole assignment is possible for state feedback and minimum order state estimator.

Pulse transfer function of the system (14) is given by

$$G_d(z) = \frac{[-0.113z^2 - 0.423z - 0.0968] \times 10^{-6}}{z^3 - 2.76z^2 + 2.52z - 0.76} \quad (15)$$

Pole-Zero map for pulse transfer function (15) is shown in figure. 4. Bode plots are shown in figure. 5.



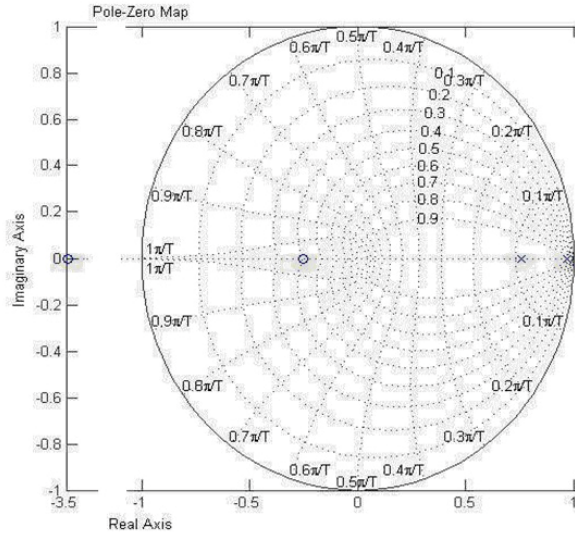


Figure 4. Pole Zero Map for the discrete system.

General expression for 3rd order pulse transfer function is given by,

$$G_d(z) = \frac{B(z)}{A(z)} = \frac{b_0 z^3 + b_1 z^2 + b_2 z + b_3}{a_0 z^3 + a_1 z^2 + a_2 z + a_3} \quad (16)$$

Comparing equations (15) and (16) we can get the values of coefficients of zero polynomial B(z) and pole polynomial A(z). Pulse transfer function in equation (15) is strictly proper with pole-excess or relative order of 1. Next section describes controller design for this system.

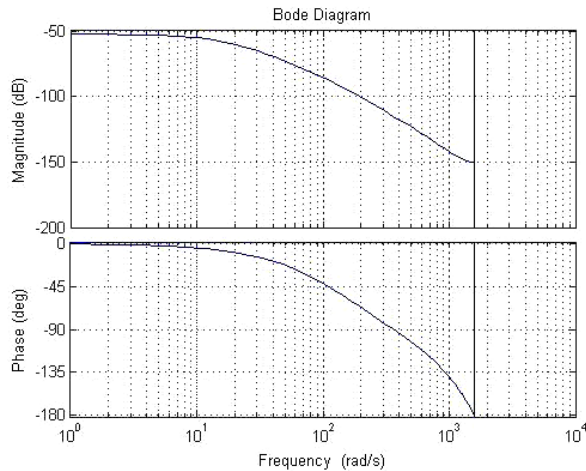


Figure 5. Bode plots of the discrete system.

4. Diophantine Estimator Design

Let $A(z)$ be a monic polynomial of order 'n' and $B(z)$ a polynomial of order 'm', where $m \leq n$, $H(z)$ be the desired characteristic polynomial for pole assignment part and $F(z)$ be the characteristic polynomial for minimum order estimator part then Diophantine Equation is given by,

$$\alpha(z)A(z) + \beta(z)B(z) = F(z)H(z) = D(z) \quad (17)$$

Where $\alpha(z)$ and $\beta(z)$ are polynomials of order 'n-1' and $D(z)$ is polynomial of order '2n-1', [1]

General control system block diagram with Diophantine estimator is shown in figure. 6. For our system $n=3$, so system general pulse function of estimator is given by following equation,

$$G_{DE}(z) = \frac{\beta(z)}{\alpha(z)} = \frac{\beta_0 z^2 + \beta_1 z + \beta_2}{\alpha_0 z^2 + \alpha_1 z + \alpha_2} \quad (18)$$

Coefficients α_i and β_i in equation (18) are to be determined. Let's put them in a vector M as in (19)

$$M = [\alpha_2 \quad \alpha_1 \quad \alpha_0 \quad \beta_2 \quad \beta_1 \quad \beta_0]^T \quad (19)$$

According to Diophantine polynomial control theory [17], M can be determined by equation (20)

$$M = E^{-1}D \quad (20)$$

Here D is vector of coefficients of polynomial $F(z)H(z)$ in equation (17). Output is the only state that is measurable so order of minimum estimate polynomial $F(z)$ is 2 resulting in the order of $H(z)$ to be 3. For dead beat response, we select all zeros of $H(z)$ and $F(z)$ at origin. So $F(z) = z^2$ and $H(z) = z^3$. It gives us $D(z) = z^5 = 0z + 0z^2 + 0z^3 + 0z^4 + 1z^5$. In vector form,

$$D = [0 \quad 0 \quad 0 \quad 0 \quad 0 \quad 1]^T \quad (21)$$

'E' is the Sylvester's matrix given by equation (22).

Using equations (15) and (16) coefficients a_i and b_i can be determined.

$$E = \begin{bmatrix} a_3 & 0 & 0 & b_3 & 0 & 0 \\ a_2 & a_3 & 0 & b_2 & b_3 & 0 \\ a_1 & a_2 & a_3 & b_1 & b_2 & b_3 \\ a_0 & a_1 & a_2 & b_0 & b_1 & b_2 \\ 0 & a_0 & a_1 & 0 & b_0 & b_1 \\ 0 & 0 & a_0 & 0 & 0 & b_0 \end{bmatrix} \quad (22)$$

Putting values of a_i and b_i in (22) we get expression for E given by equation (23),

$$E = \begin{bmatrix} -0.76 & 0 & 0 & -9.86e-8 & 0 & 0 \\ 2.52 & -0.76 & 0 & -4.32e-7 & -9.86e-8 & 0 \\ -2.76 & 2.52 & -0.76 & -1.13e-7 & -4.32e-7 & -9.86e-8 \\ 1 & -2.76 & 2.52 & 0 & -1.13e-7 & -4.32e-7 \\ 0 & 1 & -2.76 & 0 & 0 & -1.13e-7 \\ 0 & 0 & 1 & 0 & 0 & 0 \end{bmatrix} \quad (23)$$

Using equations (18) and (23) in (20) we get the value of M given by equation (24),

$$M = [0.4065 \quad 1.8752 \quad 1 \quad -3.1e6 \quad 9.4e6 \quad -7.8e6]^T \quad (24)$$

Using (18) and (24) the pole and zero polynomials of Diophantine estimator are given by (25) and (26) respectively as follows:

$$\alpha(z) = z^2 + 1.8752z + 0.40653 \quad (25)$$

$$\beta(z) = -7.82 \times 10^6 z^2 + 9.38 \times 10^6 z - 3.13 \times 10^6 \quad (26)$$



Using equations (25) and (26) in (18) we get Diophantine minimum order estimator as given by equation (27)

$$G_{DE}(z) = \frac{-7.82 \times 10^6 z^2 + 9.38 \times 10^6 z - 3.13 \times 10^6}{z^2 + 1.8752z + 0.40653} \quad (27)$$

Closed-loop simulation results: Close-loop pulse transfer function from block diagram of figure.6 is given by,

$$Y(z)/R(z) = K_o (\alpha(z)B(z))/(H(z)F(z)) \quad (28)$$

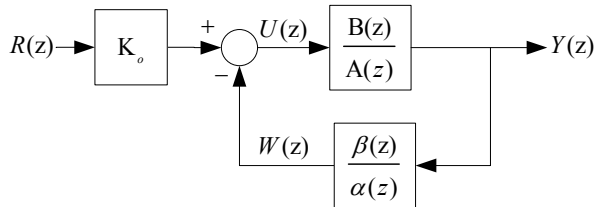


Figure 6. Block diagram of the control system.

Value of K_o can be determined by using the final value theorem for zero steady state error to a unit step input and it comes out to be,

$$K_o = \frac{H(1)F(1)}{\alpha(1)B(1)} = -4.8 \text{ e } 5 \quad (29)$$

Hence,

$$\frac{Y(z)}{R(z)} = \frac{0.054z^4 + 0.3z^3 + 0.45z^2 + 0.17z + 0.02}{z^5} \quad (30)$$

figure. 7 shows response to a unit step input. Response is clearly dead-beat and settles down within six sampling periods, equal to the order of Matrix 'E', [1].

Response to unit ramp function is shown in figure. 8 with a sampling interval of 2msec. Response to square waveform reference input is shown in figure. 9. Response to sinusoidal excitation is shown in the figure. 10. System is able to tracks sinusoidal excitation very well. Our close loop system has a bandwidth of 35Hz.

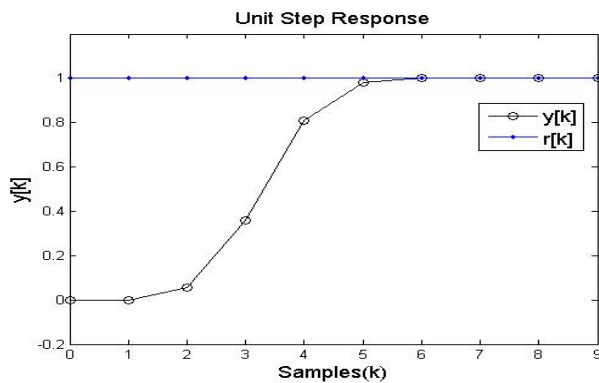


Figure 7. Unit step response of Close-loop System

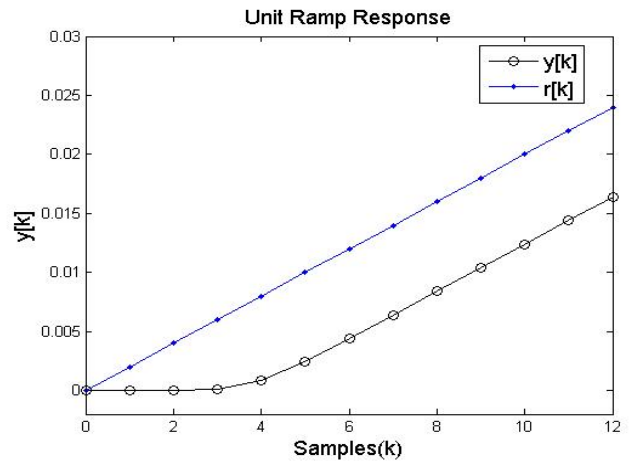


Figure 8. Unit ramp response of Close-loop System.

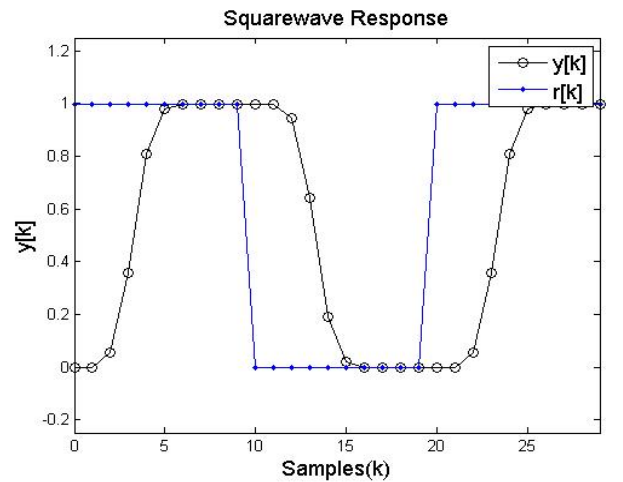


Figure 9. Square wave response of Close-loop System.

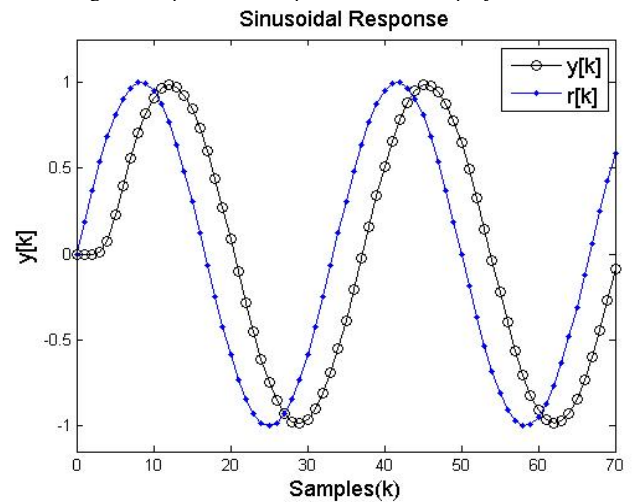


Figure 10. Sinusoidal response of Close-loop System.

Response to square input with added Gaussian white noise of 500Hz bandwidth is shown in figure. 11. System response is clearly stable against such noise.



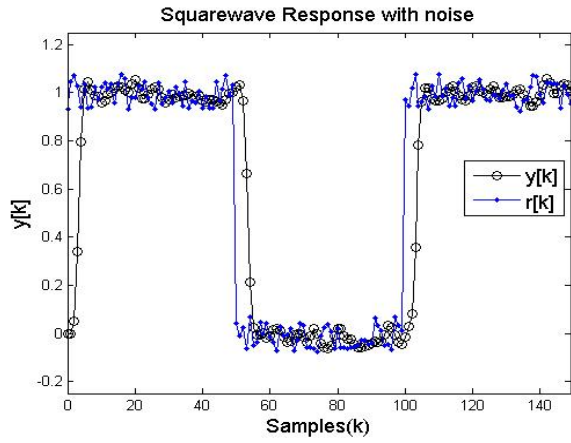


Figure 11. Square wave response with Gaussian noise.

5. Diophantine Estimator Hardware Implementation and Results

Using figure. 6 equation (27) can also be written as,

$$G_{DE}(z) = \frac{\beta(z)}{\alpha(z)} = \frac{W(z)}{Y(z)} \quad (31)$$

$$\Rightarrow \beta(z)Y(z) = \alpha(z)W(z)$$

By using equations (25) and (26), polynomial in equation (31) can be expressed by difference equation (32),

$$\begin{aligned} -7.82 \times 10^6 y(k+2) + 9.38 \times 10^6 y(k+1) \\ -3.13 \times 10^6 y(k) = w(k+2) + 1.87 w(k+1) \\ + .406 w(k) \end{aligned} \quad (32)$$

Rearranging equation (32) we get difference equation (33) which can be programmed in microcontroller.

$$\begin{aligned} w(k) = -1.87 w(k-1) - .406 w(k-2) \\ -7.82 \times 10^6 y(k) + 9.38 \times 10^6 y(k-1) \\ -3.13 \times 10^6 y(k-2) \end{aligned} \quad (33)$$

From figure. 6. control input is given by,

$$u(k) = K_o r(k) + w(k) \quad (34)$$

Using equation (33) in (34) we get,

$$\begin{aligned} u(k) = K_o r(k) - 1.87 w(k-1) - .406 w(k-2) \\ -7.82 \times 10^6 y(k) + 9.38 \times 10^6 y(k-1) \\ -3.13 \times 10^6 y(k-2) \end{aligned} \quad (35)$$

Difference equation (35) can be programmed in a digital controller. Hardware implementation strategy is diagrammatically shown in the figure. 12.

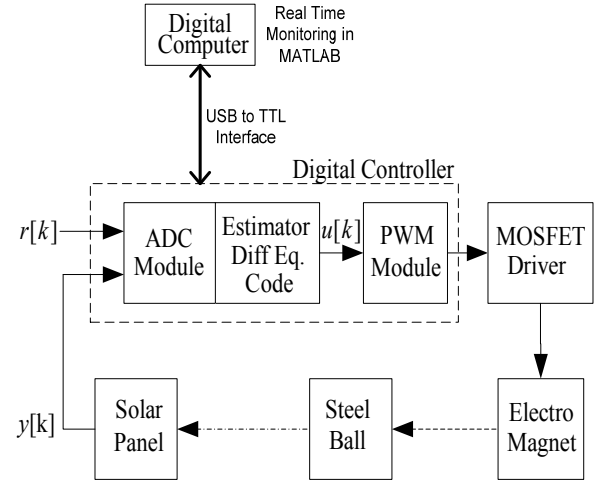


Figure 12. Control system hardware block diagram.

AVR digital controller is programmed to read reference input $r[k]$ and position of the ball $y[k]$ from solar panel using two of the on-chip ADC modules. After scaling, these variables are fed to the difference equations (33) and (35). Resulting control signal $u[k]$ is pulse width modulated. PWM output signal is applied at the gate of N-channel MOSFET driver. This FET driver controls the strength of magnetic field generated by electromagnet which in turn actuates the position of the ball.

Diophantine minimum order estimator designed in section 5 was tested on actual hardware platform of figure.1. Oscilloscope screen snapshot of the response of the system to a square wave reference input shown in figure. 13. Bottom wave form is the input and top waveform is the system response.

Oscilloscope screen snapshot of the response of the system to a square wave reference input with added white Gaussian noise of 500Hz bandwidth is shown in figure. 14. Response is clearly deadbeat and withstands noise. Oscilloscope screen snapshot of the response of the system to a sinusoidal reference input is shown in figure. 15.

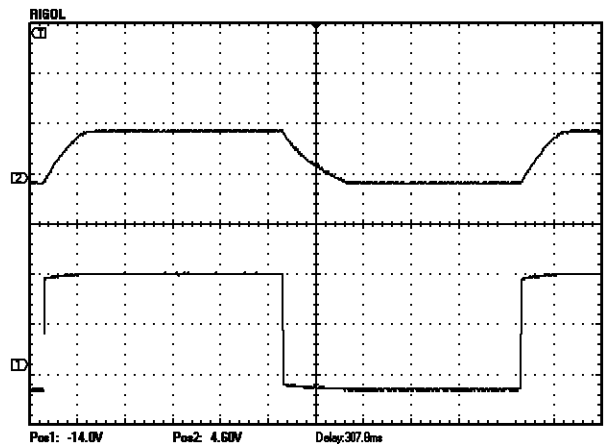


Figure 13. Square wave input response snapshot

CRO screen snapshot of the system response to a trapezoidal reference input is shown in figure. 16. The response of the system to trapezoidal reference input with added white Gaussian noise of 500Hz bandwidth is shown in figure. 17.



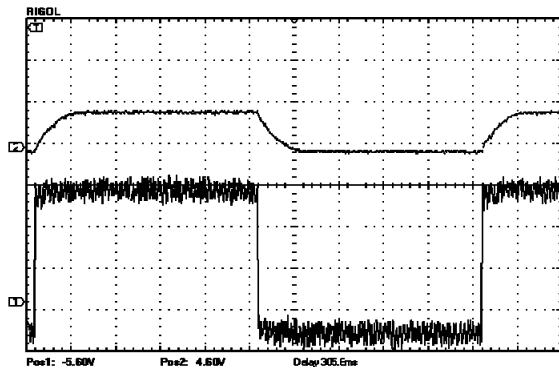


Figure 14. Square wave input with noise response snapshot

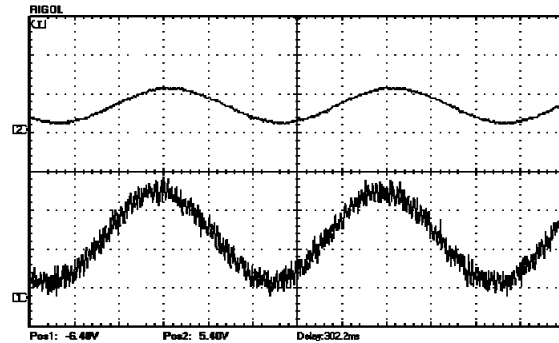


Figure 15. Sine wave input with noise response snapshot

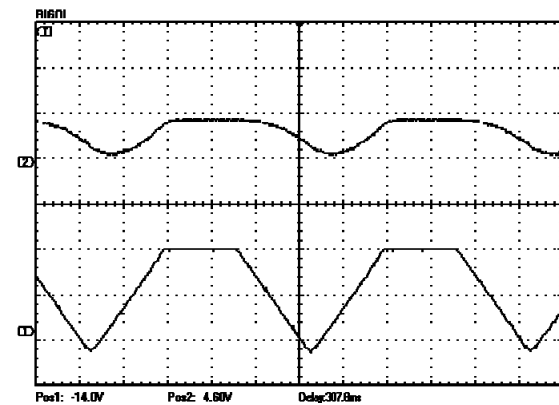


Figure 16. Trapezoidal input response snapshot

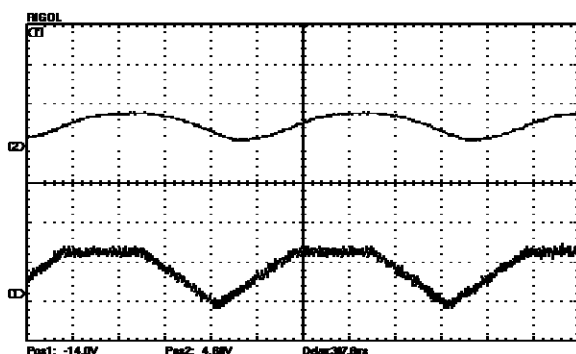


Figure 17. Trapezoidal input with noise response snapshot

6. Conclusions

If results of the classical controller design techniques are compared with the simulation and testing results presented in this paper then It can be concluded that controller designed for magnetic levitation system using Diophantine minimum order estimator approach exhibits superior response in terms of transient behavior, robustness against noises and reference tracking. Current setup involves 8bit digital ATmega controller resulting in round off and truncation errors in difference equation implementation. However these errors can significantly be improved by employing 16 bit and 32bit controllers.

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SCADA based Control of Gas Plant Inlet Pressure by Manipulating the Choke Valve of the Highest Pressure Well

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Abstract

Controlling the gas plant inlet pressure can be achieved by manipulating the choke valve existing at the more powerful remote well. The transport delay is introduced in such control system. This transport delay affects the plant inlet pressure control system stability. The main aim of this paper is to design a control methodology that overcomes this problem. The proposed strategy was simulated and compared to the classical strategies. To apply this methodology practically a communication system between the well head and the plant central control system was established. The established communication system solved the problem of the geographical nature, achieved low economic cost and the relocation possibility of the wells. The installed proposed system achieves successful communications with the remote wells. This paper presents the solution of these problems and provides the integration of the practical and theoretical part of this solution.¹

Keywords: Control, Gas plant, Inlet pressure, Communication system, Gas wells, Choke valve.

Nomenclature

PID	Proportional Integral Derivative
CCR	Central Control Room
VHF	Very High Frequency
MTU	Master Terminal Unit
RTU	Remote Terminal Unit
DCS	Distributed Control System
HMI	Human Machine Interface
WHCP	Well Head Control Panel
PLC	Programmable Logic Controller
VDC	Voltage Direct Current
VAC	Voltage Alternative Current
HPM	High Performance Manager
FTA	Field Terminal Assembly
PV	Process Value
SP	Set Point

Introduction

Natural gas is valuable both as a clean source of energy and as a chemical feedstock. Before reaching the customer, it has to pass several processing steps. These steps are partly necessary to be able to transport the gas over long distances and partly necessary for the recovery of valuable components contained in the gas. These processing steps are performed by means of gas plants such as Tarek gas plant (case study). A natural gas processing plant is a facility designed to receive raw natural gas from the remote wells and “clean” it by separating impurities and various non-methane hydrocarbons and fluids. The plant is fed from different remote gas wells. The plant inlet pressure control system is considered the most important control system in the plant. The plant inlet pressure control is required for proper operation and protection of the plant pipelines and equipments. It’s achieved by manipulating the choke valve of the more powerful gas well. To perform this remotely within the gas plant central control system, two problems raised. The first problem is the gas transport delay time which causes instability for any classical control strategy. The second problem is the need to establish a communication link to the remote well. Establishing this communication link is difficult due to the geographical nature and the economic cost.

The remainder of the paper is organized as follows: Section (1) is a survey of the previous literature, Section (2) focuses on the proposed controller, Section (3) focuses on the proposed communication system architecture, section (4) emphasizes on the practical hardware installation where the final results are demonstrated in section (5) and finally the conclusion in section (6).

1. Literature Survey

Many of the previous literature dealt with the delay problem but no one of them hit the gas plant inlet pressure control directly such as [1] which presents an advanced control algorithm but for the chemical polymerization process. The presence of delays complicates the control design of the system. However,

¹ This study has been implemented on Tarek gas plant platform at Khaldia-Apache Corporation in Egypt.



there are different approaches to model the delay such as Smith scheme and Pade approximation methods. Although Smith scheme was firstly introduced in late 1950s, it is still fundamental and basic tool for modeling systems with time-delay [2]. What makes Smith predictor so special is that it predicts outputs against time-delays. The achieved systems after prediction can be treated as delay free systems [2] (i.e. conventional design methods can be used). However, Smith predictor can be applied only to stable systems. Modified Smith can be applied to unstable systems with certain complex approximations [2]. After modeling the delay, classical PID control can be used [3]. In state- space models, state predictor is used which is similar to modified smith predictor, but it can predict also future states of the systems under consideration [4], [5] and [6]. These papers developed a novel approach for dead time compensation for nonlinear processes. The approach structure consisted of linearizing state feedback of a nonlinear system and developing smith predictor to be used in state- space to deal with systems with delay. To compensate for the dead-time linearized system, an open- loop state observer and a linear external controller have been added. Hsiao- Ping Huang, et al., presented a modified smith predictor at low frequencies with an approximate inverse of dead time in (1990) [7]. Analysis and simulation results showed that the compensator had better disturbance rejection performance than the original Smith predictor. J. J. Hench, et al., presented dampening controllers via a Riccati equation approach in (1998) [8]. The algorithm presented in this paper did not only introduce stable solution for the system but restricted the poles of the closed loop system within predefined region in the left half plane. This had an effect of dampening the closed loop system. This was accomplished by solving a damped algebraic Riccati equation and a degenerate Riccati equation. The solution to these equations was computed using numerically robust algorithms. Riccati can be expressed in format of periodic Hamiltonian system. This periodic Hamiltonian system induced two damped Riccati equations with two different solutions (symmetric and skew symmetric solutions). These two solutions were valid. They produced different closed loop eigen values and different controller gain. This increased the design flexibility by providing an alternative solution. S.I. Niculescu, Erik I. Verriest., presented a Riccati equation approach to solve delay independent stability of linear neutral systems: in (1998) [9]. This paper focused on the problem of asymptotic stability when the system has delay in the state of linear neutral systems. Sufficient conditions were given to ensure of the existence of symmetric and positive definite solutions of a continuous Riccati algebraic matrix equation coupled with a discrete Lyapunov equation. Here, we proposed a new methodology for controlling the time delayed system. Also to apply this methodology practically, a communication link between the well head and the plant central control system (CCR) has been established.

2. The Proposed Controller

The plant inlet pressure should be in range (60 Bar to 75 Bar) in order to protect the plant packages from damage

due to over or under pressure. Controlling the inlet pressure can be realized by manipulating the choke valve of the more powerful well. Fortunately in Tarek field there's a well that provide more than 70% of the gas to the plant. This well is called Rasknays#1. The remote distance between this well and the central station causes a transport delay time which affects the control system stability. Our proposed technique aims to use a combination of the smith scheme for the process dead time and extending the idea of smith to the state space representation to enable using the pole movement technique as a controller. To illustrate this methodology we need to prove analytically the effectiveness of smith scheme cancelation of the time delay effect on the system stability. This analytical evaluation is a missing point in the previous literature. The following development of Smith algorithm is based on Despande and Ash [10]. Our process model is a simple first-order plant with time delay.

$$G(S) = \frac{K}{1+\tau S} e^{-\theta S} \quad (1)$$

The model parameters; K , τ and θ were obtained experimentally for Rasknays #1's choke valve. This process can be conceptually split into delay free system dynamics (equation 2) and a pure time delay term (equation 3).

$$G_P(S) = \frac{K}{1+\tau S} \quad (2)$$

$$T_P(S) = e^{-\theta S} \quad (3)$$

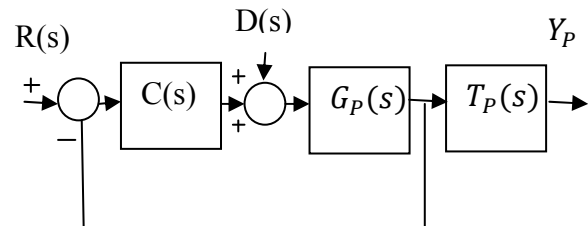


Figure 1 conventional control of time-delayed system

If the fictitious variable B figure (1) could be measured, we could connect it to the controller.

This would move the time delay outside the control loop. Since there would be no delay in the feedback signal in this case, the response of the system would be improved. Of course, this can't be done in our proposed system because our target is to control the factory inlet pressure not the well head outlet pressure; see figure 2.

Assuming $D(S)=0$ and $T_P(S) = e^{-\theta S}$; the transfer function of such system (figure 2) will be

$$\frac{Y(S)}{R(S)} = \frac{C(s)G_P(s)e^{-\theta S}}{1+C(s)G_P(s)e^{-\theta S}} \quad (4)$$

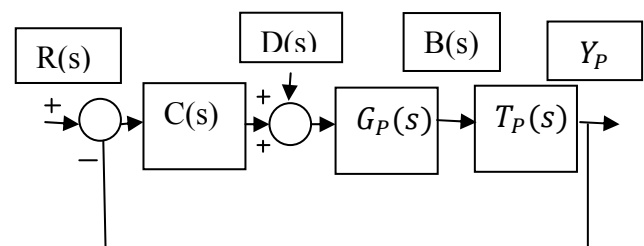


Figure 2 Typical time delayed system



From the previous equation the stability limits are not obvious. The exponential in the numerator does not bother us; therefore, it will be left undisturbed. Due to the exponential in the denominator the value of the delay θ will affect the system stability. To solve this problem the configuration of a system containing a Smith scheme is depicted in Figure 3.

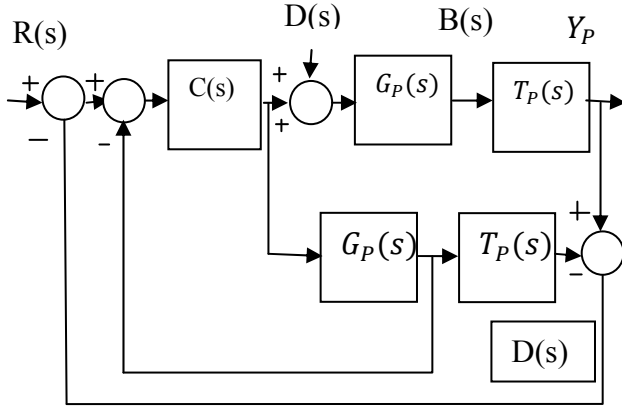


Figure 3 Complete smith scheme

Rearranging the smith predictor based on [10] will give figure 4.

The transfer function of the closed loop system relating the output control $u(s)$ to the error signal $e(s)$ is

$$\frac{Y(s)}{e(s)} = G_s(s)G(s) = \frac{C(s)G(s)}{1 + C(s)G_P(s)(1 - T_P(s))} \quad (5)$$

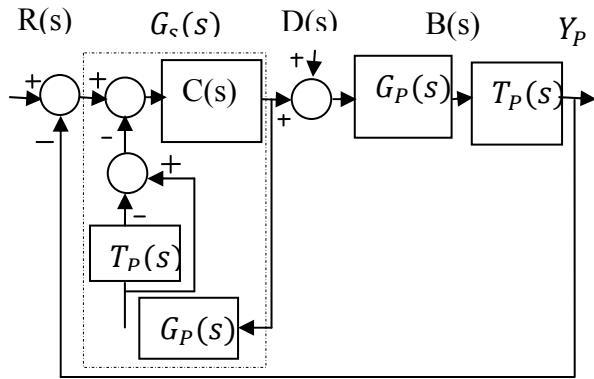


Figure 4 A rearrangement of Smith scheme

Where $C_S(s)$ is the equivalent controller; the closed loop transfer becomes

$$\frac{Y(s)}{R(s)} = \frac{C(s)G(s)}{1 + \frac{C(s)G(s)}{1 + C(s)(G_P(s) - G(s))}} \quad (6)$$

$$\frac{Y(s)}{R(s)} = \frac{C(s)G(s)}{1 + C(s)(G_P(s) - G(s)) + C(s)G(s)} \quad (7)$$

Where $G(s) = G_P(s)T_P(s)$ and $T_P(s) = e^{-s\theta}$

$$\frac{Y(s)}{R(s)} = \frac{C(s)G(s)}{1 + C(s)G_P(s)} \quad (8)$$

$$\frac{Y(s)}{R(s)} = \frac{C(s)G_P(s)T_P(s)}{1 + C(s)G_P(s)} \quad (9)$$

$$\frac{Y(s)}{R(s)} = \frac{C(s)G_P(s)}{1 + C(s)G_P(s)} e^{-\theta s} \quad (10)$$

By this modeling and using the last equation, it is clear that the stability is no longer related to the time delay as the delay has been removed from the denominator. It is clear that the delay part is shifted outside the feedback loop. So, the controller design $C(s)$ depends only on the delay free part. The previous constraints on the controller gain do not explicitly exist. This does not mean that the controller gain can take any value. The delay restricts the resultant bandwidth within certain range and therefore the gain cannot be excessively high [11]. At any case, the controller gain is to be used to compromise between the robustness and the speed of the system [12]. Due to this evaluated advantages of Smith scheme we will consider it for modeling the delay but with the presented control methodology. This presented control design methodology is theoretically to set the desired pole location and to move the pole location of the system to the desired pole location to get the desired system response. This methodology is called pole movement or placement. Actually if you tried using this strategy with a time delayed system it won't be realistic due to the effect of the term $e^{-s\theta}$ in the dominator. By means of Smith scheme this effect will be eliminated. This pole movement control method achieves the desired system response and is easy to fine the gain mathematically. To illustrate using the Smith scheme with this method we need to represent the Smith scheme for our proposed model by the state space representation. This does not mean that an alternative or different version of Smith scheme will be used. But rather, the approach is to discuss how the Smith scheme can be used with state space model. Consider a linear process without dead-time of the following state space form

$$\dot{x} = Ax + Bu(t) \quad (11)$$

$$y = Cx \quad (12)$$

$$xs = Ax + Bu(s) \quad (13)$$

$$x = \frac{B u(s)}{sI - A} \quad (14)$$

$$\frac{y(s)}{u(s)} = \frac{CB}{sI - A} \quad (15)$$

To place the system poles in the desired location to obtain the optimum performance we will insert a state feedback gain k see figure 5 where $u = v - kx$.



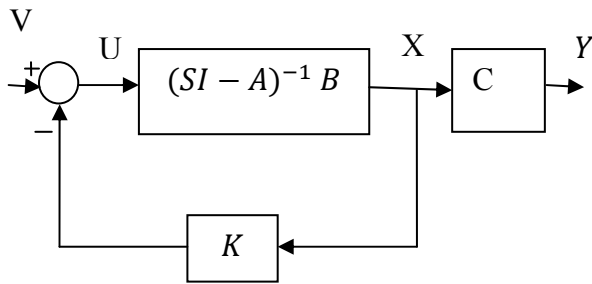


Figure 5 State feedback for a system without delay

The gain k can be chosen using algebraic riccati equation (ARE). If a time delay θ is introduced $\dot{x} = Ax + Bu(t - \theta)$ we need to use the smith scheme to eliminate the delay effect and to enable placing the poles by means of the feedback gain, see figure 6.

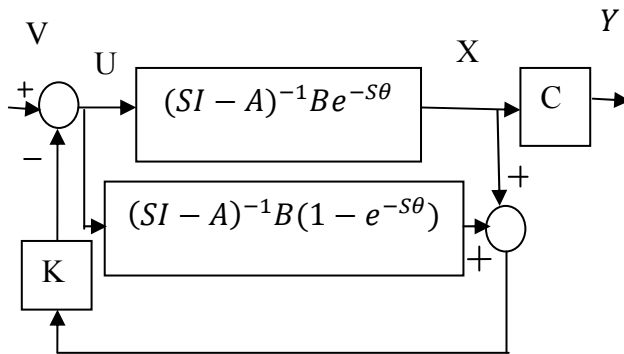


Figure 6 smith structures in state space and state feedback gain

This proposed methodology was compared to the case of using Pade approximation and also compared to the conventional PID controller with both pade and smith approximations for delay. The time response case of changing input in the inlet pressure operating range from 60 Bar to 65 Bar and also from 70 Bar to 65 Bar was shown in figure 7.

It's clear that the proposed protocol outperforms the conventional PID with smith structure and the other strategies. To apply this proposed methodology a communication system between the central plant and Rasknays#1 well head is described in the following section.

3. Stability Analysis

The Bounded Input Bounded Output (BIBO) definition of stability will be used. It states that a system is stable if the output remains bounded for all bounded (finite) inputs. Practically, this means that the system will not "blow up" while in operation. The transfer function representation is especially useful when analyzing system stability. If all poles of the transfer function (values of s at which the denominator equals zero) have negative real parts, then the system is stable. If any pole has a positive real part, then the system is unstable. If we view the poles on the complex s -plane, then all poles must be in the left half plane (LHP) to ensure the system stability. Figures 8, 9, 10 and 11 are the pole zero map for the techniques under investigation.

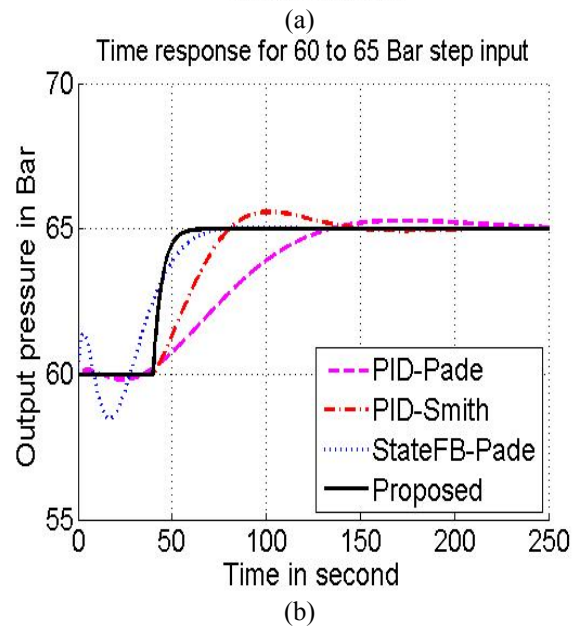
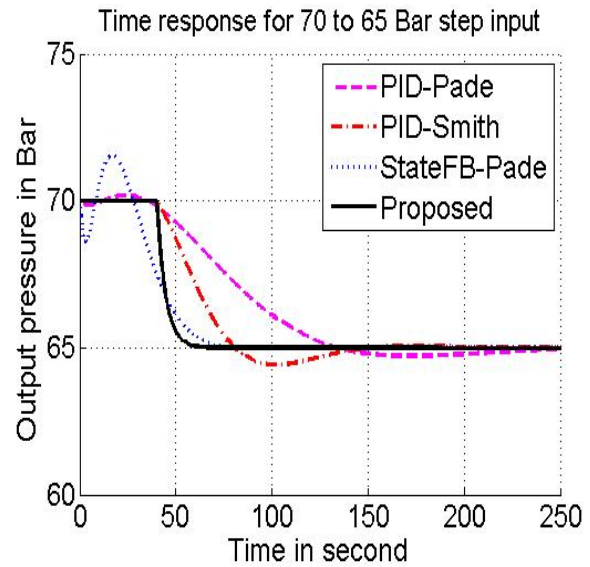


Figure 7 Time response case of (a) changing pressure from 60 to 65 Bar (b) changing pressure from 70 to 65 Bar.

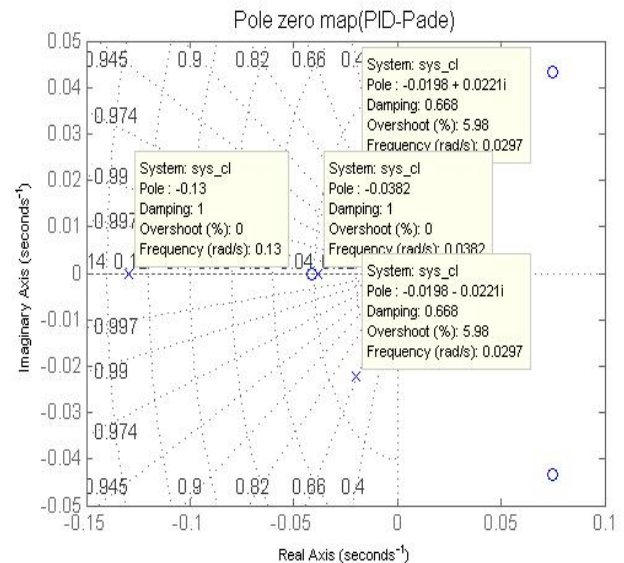


Figure 8: Pole zero map of PID-Pade



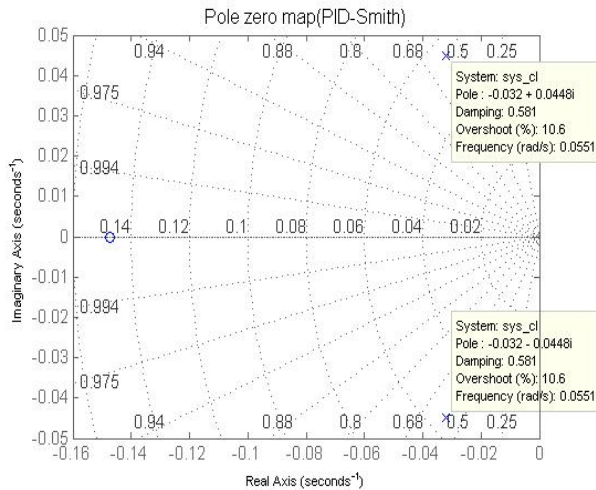


Figure 9: Pole zero map of PID-Smith

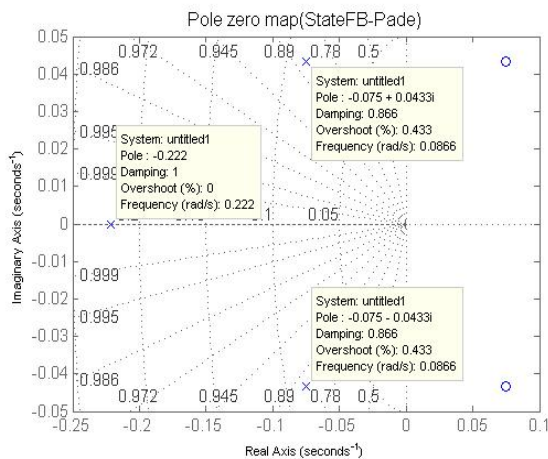


Figure 10: Pole zero map of State FB-Pade

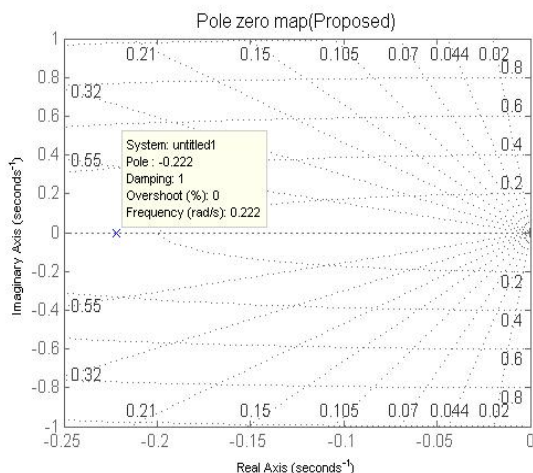


Figure 11: Pole zero map of the proposed technique

4. Communication System Architecture

The second problem is the difficulty of establishing the communication link to this remote well due to the geographical nature and the high economic cost of this project. We proposed a new communication system to overcome these difficulties Figure (12). That will be a completion of the solution proposed in [13]. The

proposed system architecture is a modified VHF system using a mid station tower. An existing telecom tower 135m altitude is used as an intermediate station between the CCR and the wells. Two antennas are installed at this telecom tower one of them is an omni directional antenna to collect the data from the directive antennas installed at the wells. The other antenna is a directive antenna to radiate towards the other directive antenna at the CCR building Figure (13). The omni-directional antenna was connected to the 25 watt modem by coaxial cable. This 25 watt modem is connected to a 5 watt modem by means of Modbus over serial RS232 connection. This 5 Watt modem is connected to a directive antenna which was directed to the CCR, figure (14). The reason behind using 25 Watt modem for communication between the mid-station tower and wells is the long distance in between. This proposal was achieved for the worst case well (Rasknays# 1, 33 Km far from the CCR) and successful communication results are obtained.

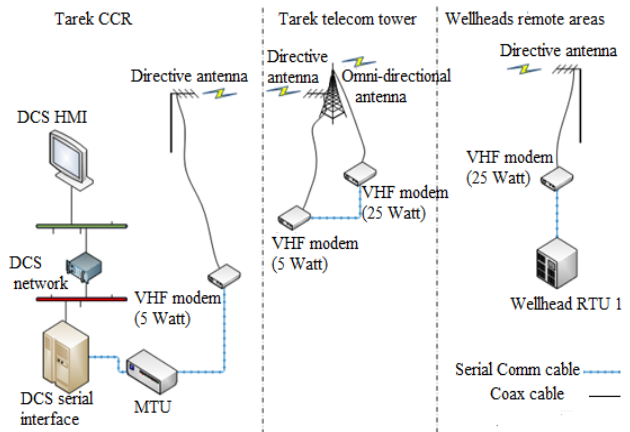


Figure 12: Structure of the proposed VHF communication system

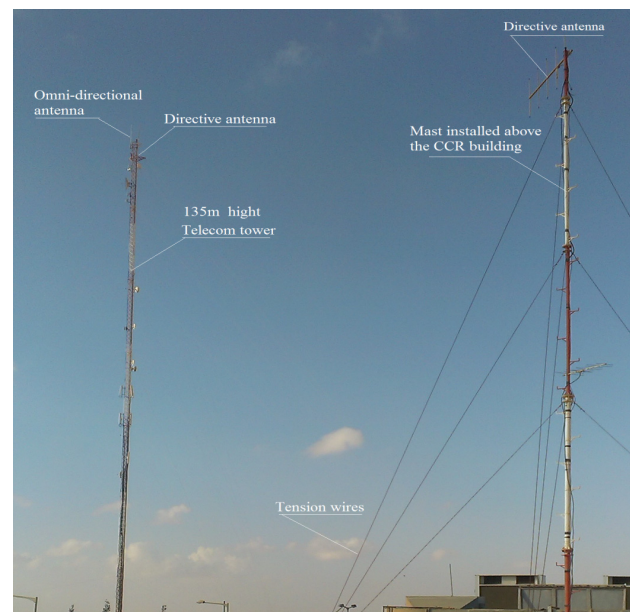


Figure 13: Two antennas (one directive and one omni-directional) at the tower and one directive antenna at CCR



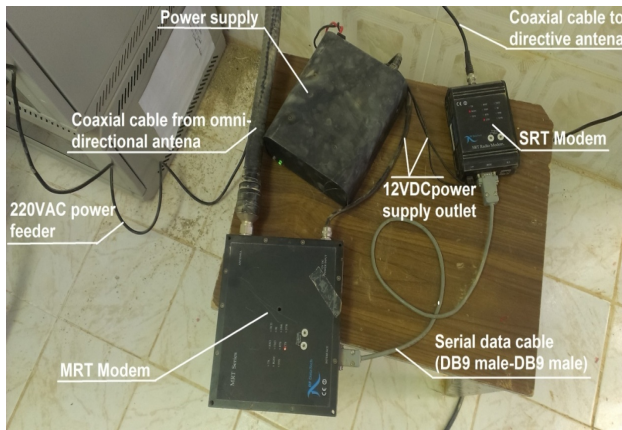


Figure 14: 25Watt modem, 5Watt modem and power supply installed at the mid-station tower

5. Hardware Equipments Installation

The system equipments were divided according to the installation place into two groups.

A. Group 1: At RTU (Rasknays1 Well head)

The well head control panel (WHCP) is a PLC acts as RTU for performing discrete control and monitoring via the HMI connected to it. This acts as a standalone controller where the well area is unsupervised area with no manpower at all. Each well has a control valve called the choke valve, three on/off valves for emergency in case of fire or any emergency conditions and many other inputs connected to the WHCP. In this project a very high frequency (VHF) communication modem was installed and connected by Modbus over serial RS232 link to the PLC of the WHCP figure (15). This modem was also connected by a coaxial cable to a directive antenna. This antenna was directed to the telecom tower and fixed at the mast existing at the well head; figure (16). The modem installed here is a 25 watt modem to cover the long distance between the well head and the tower.

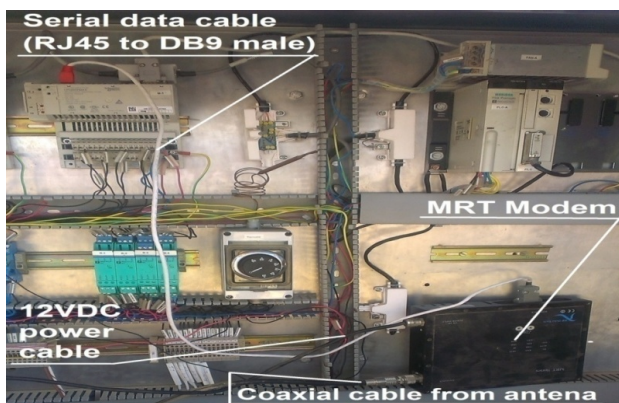


Figure 15: The VHF 25 watt modem connected to the RTU

At CCR, A directive antenna was installed and directed towards the tower and connected by a coaxial cable to a 5watt modem. This modem was connected to the MTU by means of Modbus over RS232 serial cable. When the MTU asks the desired information the RTU sends it so the MTU is considered as master and RTU is slave. After receiving the required data, MTU communicates to Honeywell distributed control system (Honeywell

DCS) and hence the human to machine interfaces (HMI's) through Modbus protocol over serial link. At this level of communication the MTU will be slave whereas DCS will be the master. The MTU in this project is a Schnieder PLC programmed to collect data from wells and send it via a serial link to the DCS field terminal assembly (FTA) serial interface card and then to serial interface of the high process manager (HPM). The DCS HPM is the processor which performs the control logic operations, figure. (17- 19).

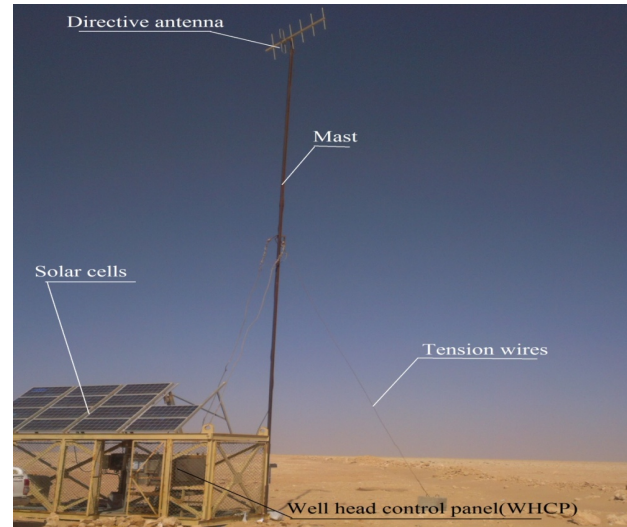


Figure 16: Directive antenna at Rasknays1 well head

B. Group 2: At MTU (CCR)

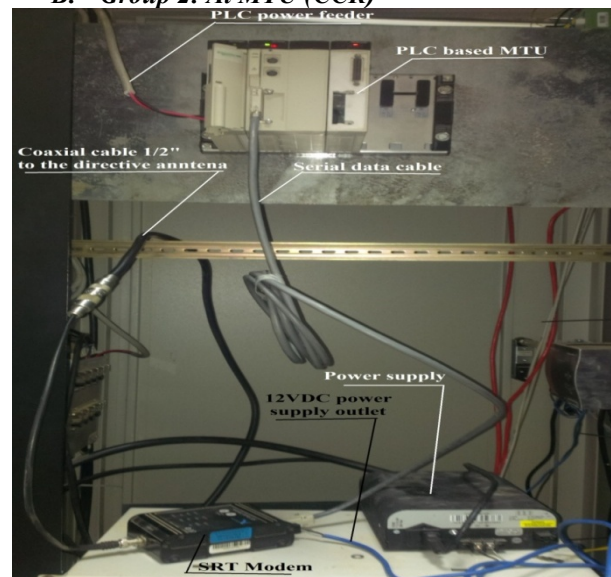


Figure 17: devices installed in CCR including MTU and 5 watt modem.

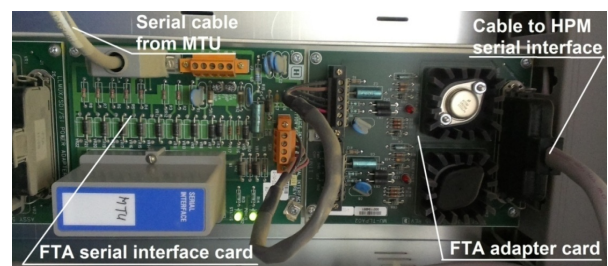


Figure 18: DCS FTA serial interface connection





Figure 19: HPM serial interface card

6. Final Results

The DCS is a Honeywell solution called TDC3000. It compares the plant inlet pressure (current value) to the required set point and calculates the choke opening percentage of Rasknays#1 choke valve according to the proposed control strategy. This computed value will be sent to the MTU and then to the RTU at Rasknays#1. In our application the RTU also is represented by Schnieder PLC. It receives the choke opening from the MTU and operates the choke valve accordingly. Beside this control job the RTU has other activities. It collects the well digital data like the fire sensor reading, the emergency shutdown button, flow line pressure high and low switches and performing emergency shutdown accordingly. It performs this emergency shutdown by closing the shutdown valves connected to the RTU output like the master, wing and subsurface on/off valves. It also collects some analog data like the flow line pressure and temperature to be monitored only. Figure (20) displays the readings of Rasknays #1 well at the DCS HMI as a result for our successful communication link.

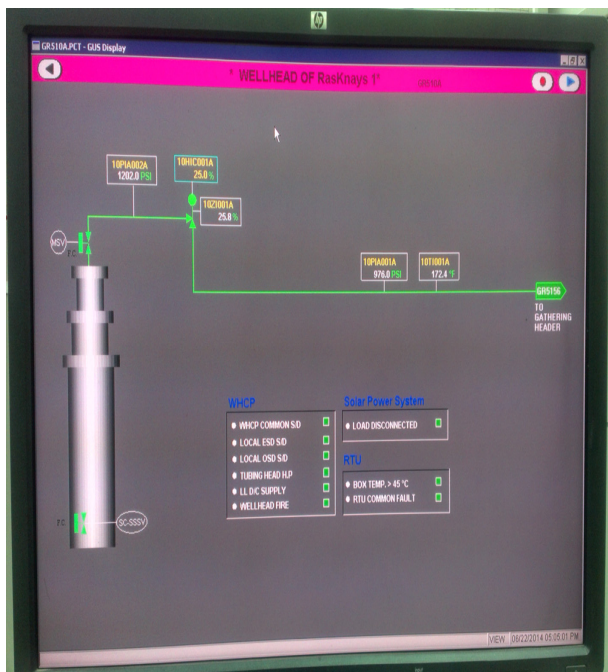


Figure 20: Data of Rasknays#1 was monitored by DCS HMI's in CCR

The result shows that the wellhead and flow line pressures are 1202 and 976 Psi respectively. And the flow line temperature is 172 F. The well valves and the digital signals are green color which means open for valves and normal condition for the digital signals. If this colors was red that will mean valve closed and up-normal digital signals. These colors were configured from the DCS HMI program. The efforts undergone in the theoretical and practical parts were combined and utilized in the inlet pressure automatic control by manipulating the choke valve of the more powerful well remotely. The trend of the pressure control loop is shown in figure 21. It should be noted that the values of the pressure process value (PV) is nearly the same as the required set point (SP) whereas the range of both is different in the vertical axis of the graph. The reason behind considering different ranges is to enable showing both curves only.

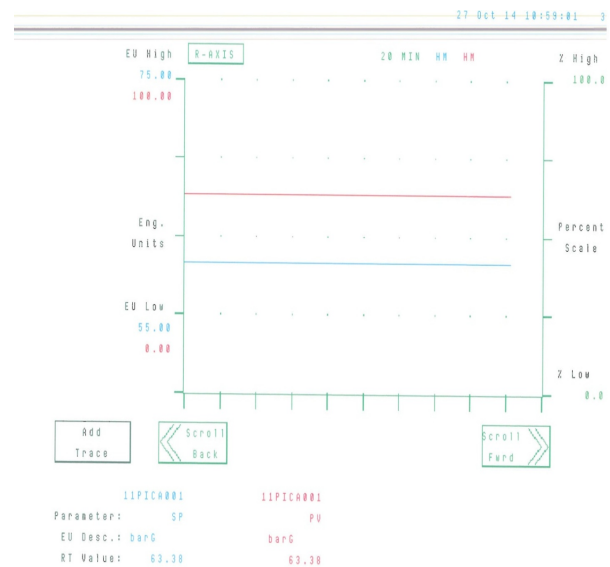


Figure 21: Inlet pressure control trend

7. Conclusions

This paper discusses the problem of transportation delay introduced in the gas plant inlet pressure control process. It shows the effect of this delay on the system stability. It proposes a new control methodology based on a hybrid of pole movement and smith scheme. The proposed techniques was evaluated and compared to the conventional PID controller with Smith scheme. The obtained simulation results show that the proposed technique outperforms the conventional PID in terms of the time response and stability metrics. The proposed control methodology was applied practically to control the gas plant inlet pressure by manipulating the choke valve of the more powerful well. For this practical application it was essential to establish a communication link between the CCR and this well. This communication link was established using VHF due to its low economic cost. It was difficult before to establish this link due to the geographic nature of this area where no line of sight between the wells and the CCR. We solved this last problem by utilizing the existing high Telecom tower. This tower was used as an intermediate station to collect the data from all wells using omni-directional antenna then sending it to the CCR.



8. Acknowledgements

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Multilevel Inverter fed Permanent Synchronous Motor with Field Oriented Control Using Three Level Space Vector Modulation

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Abstract

The necessity of increasing power quality in the past years leads to the development of inverter due to its efficiency and control methods. This paper focuses on a novel method of diode clamped multilevel inverter(DCMLI) fed permanent magnet synchronous motor(PMSM) with field oriented control (FOC) using space vector modulation(SVM). Recent advances in the power-handling capabilities of static switch devices with higher voltage rating have made the use of voltage inverter (VSI) feasible for high-power applications. High-power and high-voltage conversion systems have become very important issue for the power electronic industry handling the large ac drive and electrical power applications. MLIs have become very popular with sophisticated control techniques. Both SPWM and SVM control strategies are implemented to obtain multilevel output voltages. In this work, DCMLIs for PMSM with FOC is carried out in MATLAB/ Simulink/ Sim Power System environment. Comparative analysis is investigated for total harmonic distortion (THD) using both control strategies up to seven with SVM and nine levels with SPWM

Keywords: SPWM, SVM, PMSM and FOC

Nomenclature

V_d	direct axis stator voltage, V
V_q	quadrature axis stator voltage, V
i_d	direct axis stator current, A
i_q	quadrature axis stator current, A
R_s	Stator armature resistance, Ω
p	no. of poles
L_d	direct axis inductances, H
L_q	quadrature axis inductances, H
λ_d	direct axis flux linkage, wb
λ_q	quadrature axis flux linkage, wb
λ_f	magnetic flux linkage, wb
ω_e	rotor speed in electrical, rad/s
ω_m	mechanical speed
J	moment of inertia, kg.m ²
B	friction vicious gain, Nm/rad/s
T_e	electromagnetic torque, Nm
T_L	load torque, Nm

Acronyms

DCMLI	Diode Clamped Multi Level Inverter
PMSM	Permanent Magnet Synchronous Motor
FOC	Field Oriented Control
SVM	Space Vector Modulation
SPWM	Sinusoidal Pulse Width Modulation

1. Introduction

There are generally three methods for higher inverter output voltage; connecting devices in series, using switching devices with higher voltage rating, or adopting multilevel topology. The first method can be used only when the problem of dynamic voltage sharing has been efficiently solved. Otherwise it should be avoided. The choice of the second method is really a complicated compromise of the cost, reliability and availability. Given limited device voltage ratings, the most effective solution to higher inverter voltage is multilevel topology. The theory of multilevel converters has been implemented from 1975. Multilevel started with the three-level converter. Consequently, a number of multilevel converter topologies have been developed.

The proposed three-level diode clamping inverter has an advantage in high speed drive or active power filtering areas where high switching frequency operation is necessary [1]. In order to improve the quality of an inverter the power conversion is performed in the form of small voltage steps resulting in lower harmonics [2]. In high power applications, the multilevel inverters are more adequate compared to the conventional two-level structure [3]. The most popular method of controlling the output voltage is termed as pulse width modulation technique PWM is an internal control method and it gives better result than an external control method.

A three-level neutral-point-clamped inverter using triangular carrier pulse width modulation is presented in [4] and [5]. This technique is suitable for reactive power compensation and in FACTS application where inverters operation with a relatively low switching frequency required [6]. Also, using the multi carrier pulse width modulation (MCPWM) technique, good performance can be obtained for high power for applications [7]. Recently



Japan electric train corporation have developed a three level PWM inverter for traction drives [8].

In this paper the performance of FOC of PMSM DCMLIs is investigated with DCMLIs using SPWM and SVM control strategies. The reason to select PMSM is due to its high efficiency, high torque-to-inertia ratio, high power factor, fast response and rugged construction less audible noise and low maintenance cost [9]. [10]. These features make the extensive usage of PMSM in applications like electrical vehicles, aircraft, nuclear power station ship engines, robotic automation, escalators and industrial servo drives [11].

The paper is organized as follows: Section II revolves around the introduction of multilevel inverters. Section III explains control strategies SPWM and SVM. Section IV deals with the modelling of PMSM. Section V focuses on the simulation results of different levels of phase and line voltages and the current waveforms of the DCMLI. It also gives a comparative analysis of THD, speed and torque responses of PMSM. Section VI summarises the conclusions drawn from the work with future recommendations.

2. Multilevel Inverters

There are generally three methods for higher inverter output voltage; connecting devices in series, using switching devices with higher voltage rating, or adopting multilevel topology. The first method can be used only when the problem of dynamic voltage sharing has been efficiently solved. Otherwise it should be avoided. The choice of the second method is really a complicated compromise of the cost, reliability and availability. Given limited device voltage ratings, the most effective solution to higher inverter voltage is multilevel topologies.

One of the multilevel structures that has gained much attention and widely used is the Neutral-Point-Clamped multilevel inverter or also known as Diode Clamped multilevel inverter. This structure was first proposed in 1980 [12]. Basically, NPC multilevel inverters synthesize the small step of staircase output voltage from several levels of DC capacitor voltages. An m -level NPC inverter consists of $(m-1)$ capacitors on the DC bus, $2(m-1)$ switching devices per phase and $2(m-2)$ clamping diodes per phase.

The DC bus voltage is split into 3 levels by using 2 DC capacitors, C_1 and C_2 . Each capacitor has $V_{dc}/2$ volts and each voltage stress will be limited to one capacitor level through clamping diodes [13].

The number of levels can be extended to a higher level by additional switching devices and with these additions, the inverter will be able to achieve higher AC voltage, producing more voltage steps that will be approaching sinusoidal with minimum harmonics distortion. During inverter operations, the switches near the centre tap are switched on for a longer period compared to the switches further away from the centre tap. As the switch is further away from the centre tap the switching time is shorter. Another difference between the conventional 2-level and multilevel NPC is the clamping diode. In case of 3-level NPC inverter, clamping diode, D_1 and D_4 clamped the DC bus voltage into three voltage level, $+V_{dc}/2$, 0 and $-V_{dc}/2$.

A multilevel converter has several advantages over a conventional two-level converter that uses high switching frequency pulse width modulation (PWM). The attractive features of multilevel inverters can be briefly summarized as follows: i). MLIs produce output voltage with less distortion, dv/dt stresses hence reduce electromagnetic problems. ii). It generates small common mode voltage so that the stress in the bearings of a motor reduces to a significant extent. Also this common mode voltage can be eliminated by adopting advanced modulation strategies. iii). It draws input current with low distortion factor iv). Operation with both fundamental switching frequency and high switching frequency gives less switching loss and higher efficiency [14].

3. Control Strategies

Abundant modulation techniques and control paradigms have been developed for multilevel converters such as sinusoidal pulse width modulation (SPWM), selective harmonic elimination (SHE-PWM), space vector modulation (SVM), and others. The proposed DCPMIs are developed up to nine levels with SPWM and SVM.

A. Sinusoidal Pulse Width Modulation

Multilevel inverter topologies with sinusoidal pulse width modulation have drawn tremendous attention in the power industry since it provides the high-power. Pulse Width Modulation refers to a method of carrying information on a train of pulses, the information being encoded in the width of the pulses. The pulses have constant amplitude but their duration varies in direct proportion to the amplitude of analog signal. The following are the different phase relationships in SPWM strategies which are implemented in this work.

- i). Alternate phase disposition (APOD) – every carrier waveform is in out of phase with its neighbor carrier by 180.
- ii). Phase opposition disposition (POD) – All carrier waveforms above zero reference are in phase and are 180 degree out of phase with those below zero
- iii). Phase disposition (PD) – All carrier waveforms are in phase

B. Space Vector Modulation

The second important control strategy, space vector modulation gives fast dynamic response and wide linear range of fundamental voltage compared with the conventional pulse width modulation techniques [14]. The proposed work is based on SVM for multilevel inverter.

The SVPWM for multilevel inverters involves mapping of the outer sectors to an inner sub-hexagon sector, will be very complex, as a large number of sectors and inverter vectors are involved. [15]. Conventional space vector PWM (SVPWM) is a widely used PWM strategy which has the advantages of low harmonic distortion in the output current and suitable for digital implementation. [16]. In SVPWM technique, actual switching times can be produced by the recombination of effective voltage vectors using the information of the reference voltage vectors location. [16]. In this work, three level svm is



considered as a basis for generating 5, 7 and 9 levels. The space vector hexagon for three level is shown in figure 1. In general SVM for 5, 7 and 9 level voltage output waveform requires 125, 343 and 729 switching combinations respectively which make the control logic more complicated. In the proposed work, SVM for 5, 7 and 9 level voltage output waveform requires 54, 108 and 216 switching combinations respectively (Table 1). The flow chart for three level svm to seven level DCMLI is depicted in figure 2.

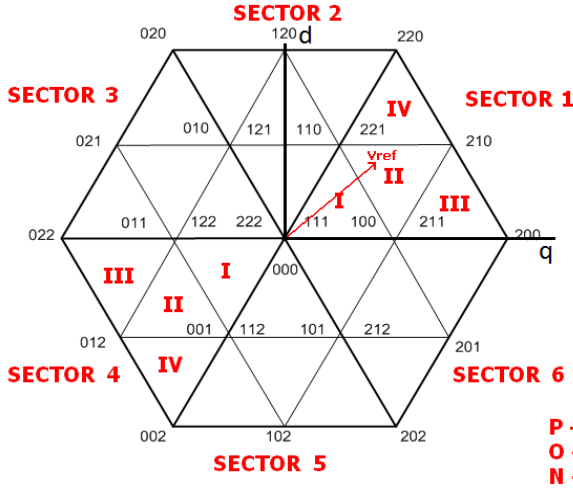


Figure 1. Space vector hexagon for three level SVM

4. Field Oriented Control of PMSM

The control methods used for the permanent magnet synchronous motors are: V/f control, field oriented control and direct torque control [17]. Field oriented control is one of the control methods of the PMSM, which achieve fast dynamic response and the flexibility normally obtained in control of separately excited dc motor. In the proposed work, DCMLI fed field oriented control of PMSM with three level space vector modulation is considered. In this work PMSM is modeled in MATLAB/SIMULINK environment. Modeling equations for PMSM are given in equations (1) to (8). The proposed block diagram of the work is given in figure 3.

$$V_d = R_s i_d + p \lambda_d - \omega \lambda_q \quad (1)$$

$$V_q = R_s i_q + p \lambda_q + \omega \lambda_d \quad (2)$$

$$\lambda_d = L_d i_d + \lambda_f \quad (3)$$

$$\lambda_q = L_q i_q \quad (4)$$

$$V_d = R_s i_d + L_d \frac{d}{dt} i_d - L_q \omega_e \frac{d}{dt} i_q \quad (5)$$

$$V_q = R_s i_q + L_q \frac{d}{dt} i_q + L_d \omega_e \frac{d}{dt} i_d + \omega_e \lambda_f \quad (6)$$

$$T_e = \frac{3}{2} p [(L_d - L_q) i_d i_q + \lambda_f i_q] \quad (7)$$

$$J \frac{d\omega_m}{dt} + B \omega_m + T_L = T_e \quad (8)$$

Generally vector control techniques are referred as the field oriented control [FOC]. The basic idea of the FOC algorithm is to resolve stator current in to field producing component and torque producing component. This is achieved easily in FOC by adapting $i_d = 0$. Then the motor control is as simple as that of a separately excited DC motor. In the proposed work, a novel DCMLI fed FOC of PMSM is implemented with SVM.

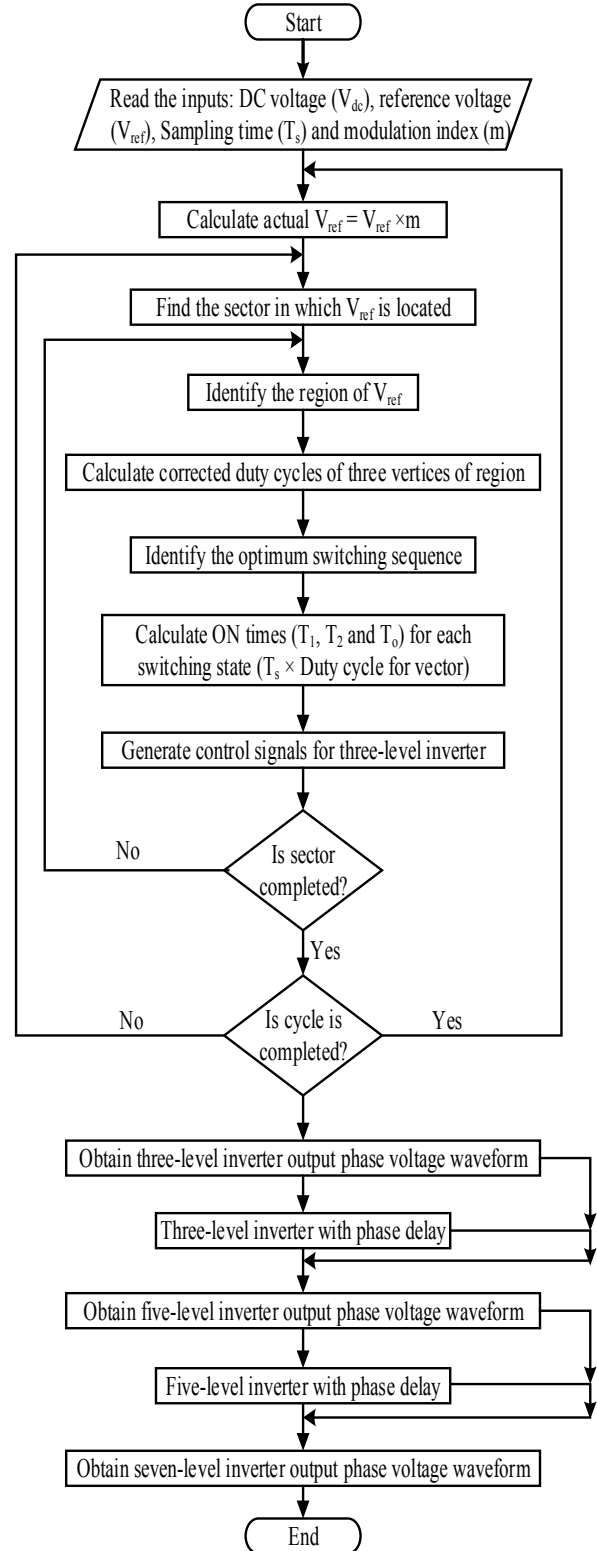


Figure 2. The flow chart for three level svm to seven level DCMLI



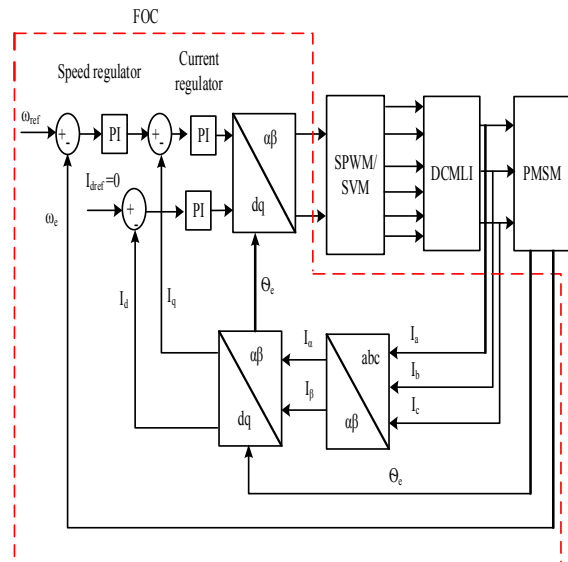


Figure 3. The proposed block diagram

5. Results and Discussion

The dynamic performance of field oriented control permanent magnet synchronous motor using three level space vector control is carried out using MATLAB/Simulink/Sim Power System environment.

There are four test cases; at first the analysis of DCMLI is carried out for RL load using SPWM. In second case the simulation is carried out for the three level, five-level, seven level and nine-level DCMLI using for PMSM with SPWM. In third case the simulation is carried out for the three level, five-level and seven level DCMLI for PMSM with SVM. Finally the dynamic analysis is carried out for FOC of PMSM with three level SVM. The results are plotted from figure 4 to figure 12.

The first case is implemented for three different phase relations of SPWM (PD, POD and APOD). The last three cases are tested under constant speed and load torque. The load torque 20Nm is applied at $t=1$ sec.

The results depict that as the level of phase voltage increases the THD decreases for phase voltage and currents. Also the torque ripples reduces with multi level phase voltages, which is clearly shown from figure 5 to figure 8 with SPWM; from figure 9 to figure 11 with SVM and from figure 12 with FOC with three level space vector modulation. In these three cases FOC with SVM gives better results with reduced harmonics, torque ripples and steady state error also got reduced to low value. The THD analysis is given in Tables 2 to Tables 6. It is very well clear that THD for phase voltage and current reduces from 37% to 3.7% and from 12% to 2.5% respectively. The THD is reduces as the level of the DCMLI increases. The input current drawn by PMSM is less distorted as level of inverter increases as well as FOC with SVM. The results have been presented and analyzed in this paper. The parameters of the PMSM and dc voltage considered for this paper is presented in Table 7.

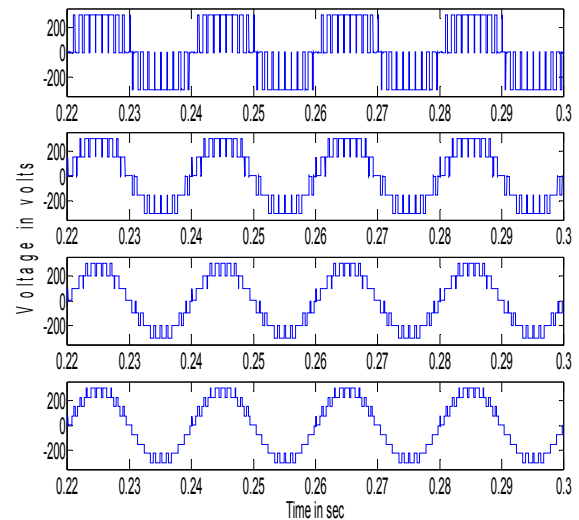


Figure 4 Phase voltage wave forms for 3, 5, 7 and 9 level DCMLI using SPWM with RL-load

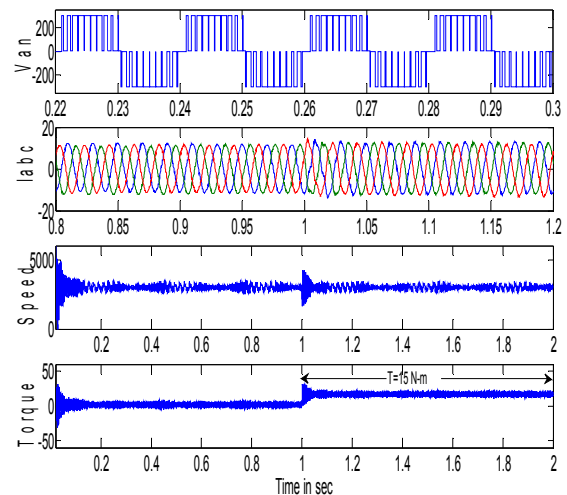


Figure 5 Phase voltage , current, speed and torque response of a PMSM 3 level DCMLI using. SPWM

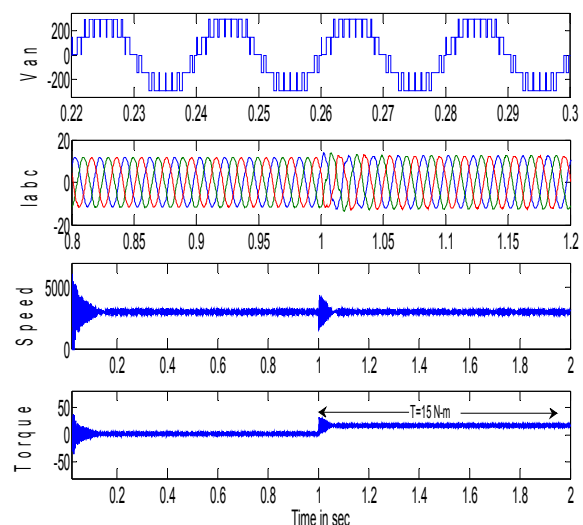


Figure 6. Phase voltage , current, speed and torque response of a PMSM 5level DCMLI using. SPWM



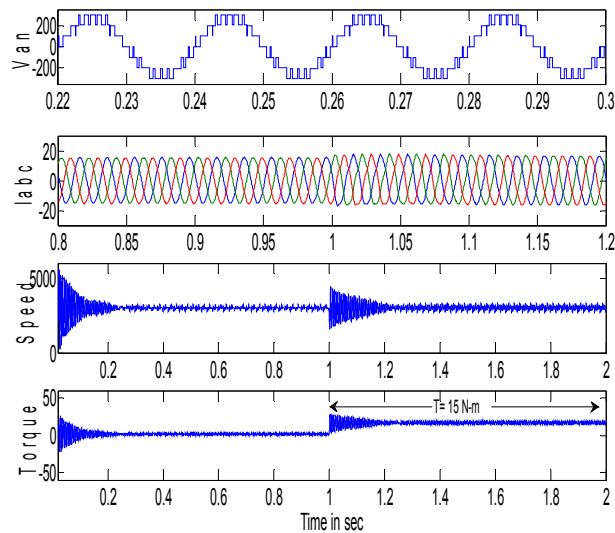


Figure7. Phase voltage , current, speed and torque response of a PMSM 7level DCMLI using. SPWM

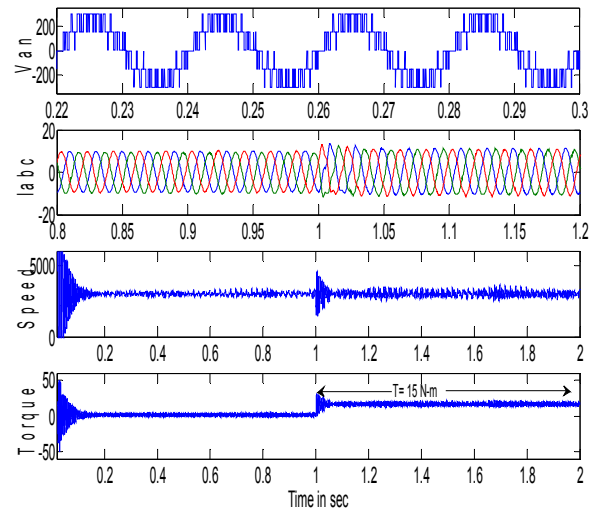


Figure 10. Phase voltage , current, speed and torque response of a PMSM 5level DCMLI using. SVM

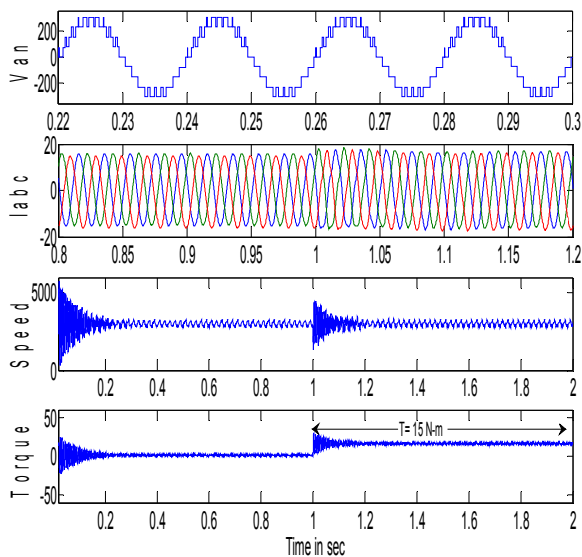


Figure 8. Phase voltage , current, speed and torque response of a PMSM 9level DCMLI using. SPWM

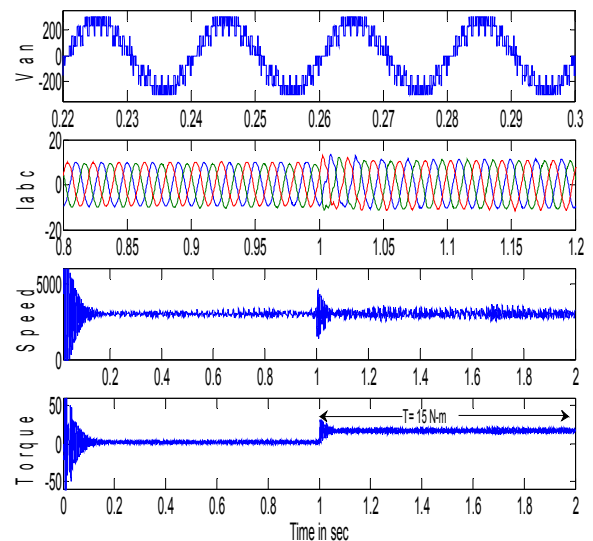


Figure 11. Phase voltage , current, speed and torque response of a PMSM 7level DCMLI using. SVM

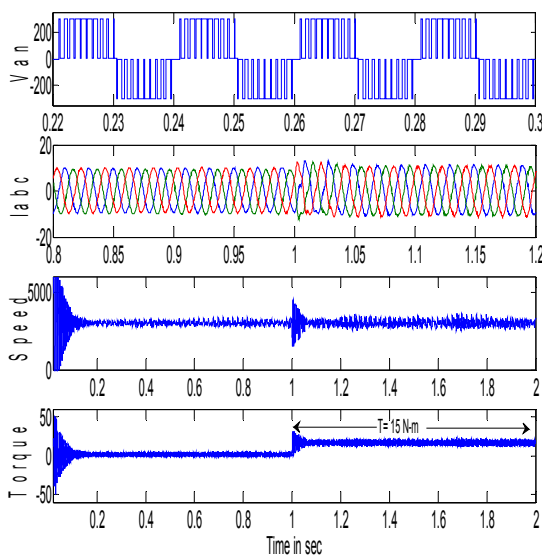


Figure 9. Phase voltage , current, speed and torque response of a PMSM 3level DCMLI using. SVM

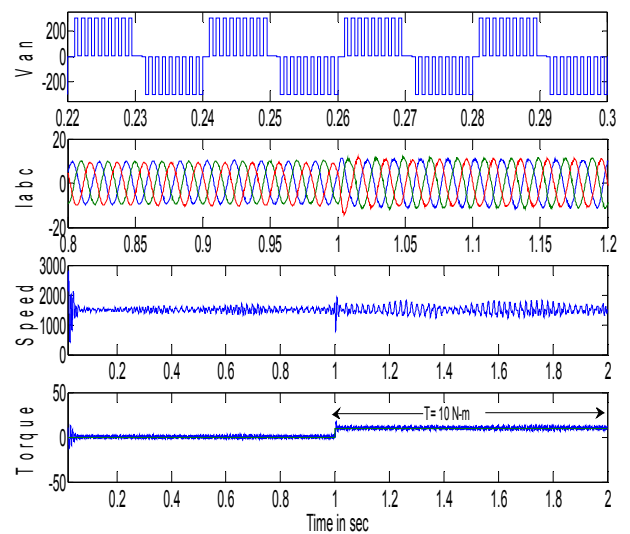


Figure 12. Phase voltage , current, speed and torque response of a FOC- PMSM 3level DCMLI using. SVM



Table 1: No of switching states required

DCMLI	SPWM	SVM	Switching states	
			Actual	Proposed
3-level	12	12	27	27
5-level	24	24	125	54
7-level	36	48	343	108

Table 2: THD analysis of phase voltage up to 9 level with RL Load

S.No	SPWM with RL-load		
	PD	POD	APOD
3-Level			
1.	37.88%	29.31%	
5-Level			
2.	23.9%	23.87%	22.89%
7-Level			
3.	18.95%	18.64%	18.38%
9-level			
4.	15.13%	15.60%	15.58%

Table 3: THD analysis of phase current up to 9 level with RL Load

S.No	SPWM with RL-load		
	PD	POD	APOD
3-Level			
1.	12.32%	11.41%	
5-Level			
2.	11.28%	10.87%	10..83%
7-Level			
3.	8.12%	7.96%	7.54%
9-level			
4.	6.82%	6.91%	5..38%

Table 4: THD analysis up to 9 level with PMSM

S.No	Levels of inverter	SPWM with PMSM	
		V _{THD}	I _{THD}
1.	3-level	37.61%	4.27%
2.	5-level	19.27%	3.74%
3.	7-level	16.07%	2.78%
4.	9-level	12.46%	3.11%

Table 5: THD analysis up to 7level using SVM

S.No	Levels of Inverter	SVM with PMSM	
		V _{THD}	I _{THD}
1.	3-level	3.77%	3.05%
2.	5-level	3.62%	3.03%
3.	7-level	3.48%	3.00%

Table 6: THD analysis for FOC of PMSM with SVM

S.No	Levels of Inverter	SVM with PMSM and FOC	
		V _{THD}	I _{THD}
1.	3-level	3.77%	2.53%

Table 7: PMSM Parameters

Parameter Description	Value
Stator Resistance	4.3 Ω
Direct Axis Inductance	0.067 H
Quadrature Axis Inductance	0.026 H
Rotor Flux Constant	0.1546 V/rad/s
Moment of Inertia	0.000179 Kg m^2
Friction Vicious Gain	0.01 Nm/rad/s
Number of Poles	4
DC Voltage	600 V

6. Conclusion

As the rated inverter output voltage is proportional to the number of active switches and the gate signals generation with SVM gives low harmonic distortion, in the this paper multi level with SVM is implemented for FOC of PMSM. The results prove that the harmonic distortion is less with SVM than SPWM. Also as the levels of output voltage increases the harmonic distortion further deceases. The torque response of a PMSM is better for 7 level diode clamped inverter. In the proposed work the switching combinations are very less as compared to the conventional five and seven level SVM algorithms. It is further suggested that the proposed work further can be implemented with fuzzy logic and model reference adaptive control techniques.

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Advances in Scheduling Approaches in Real-Time Systems

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Abstract

This study deals with various scheduling approaches practiced in real-time systems. It summarizes the advances that have occurred in scheduling approaches, ranging from conventional scheduling to the current trends including adaptive scheduling. The recent advances in handling complex multi-parameter criticality are identified and reviewed. Various challenges in transforming conventional scheduling into intelligent scheduling and the mapping of catastrophe theory into intelligent systems are some of the important aspects discussed here. The paper integrates various technical points related to a spectrum of design approaches and issues pertaining to the design as well as development of catastrophe handling process in real-time systems used in various plants.

Keywords: *catastrophe, flexible, adaptable, intelligent, scheduling, real-time systems.*

1. Introduction

A real-time system (RTS) is a system in which the response to externally generated inducement occurs within a stipulated time frame. That is; the accuracy of the response generated is determined not just by the logical result but also on the response time. Examples include spacecraft, air traffic control systems, and nuclear power plants. Such systems have a set of identifiable pieces that comprise their job that serves as a unit of work called a task. Each task is composed of five temporal parameters (A_i , R_i , C_i , Res_i , D_i). The arrival of a task at the system happens at a time instant called arrival time (A_i) and were ready for execution at its release time (R_i). The amount of time taken for its execution is called computation time (C_i), and the task finishes its execution at its response time (Res_i). The execution of a task must be completed by its deadline (D_i). Failure to meet the deadline may lead to disaster in hard real-time systems whereas missing a deadline causes only a certain amount of performance degradation in soft real-time system. Each task must satisfy the condition which states that the summation of the release time and the computation time must be

equal to the response time, which is required to be less than or equal to the deadline, as given below in Equation 1.

$$R_i + C_i = Res_i \leq D_i \quad (1)$$

This paper reviews the implementation methods and practices prevailing in real-time system design and implementation, from its inception to the current period. In 1952, such activities were called "real time computer programs." In those days, only the basic characteristics of the task, such as release time and computation time, were addressed in the real-time system design methods for information processing systems. These parameters are insufficient to satisfy the requirements of real-time system... Every task is expected to achieve completion before a time instant called the deadline that determines the worst-case time that could be taken for executing that task. The concept of the deadline has been examined using "Simulation techniques for the testing and evaluation of real-time programs and their distinctive characteristics" [1]. The use of large-scale systems requires more attention to the areas of testing and evaluation of real-time programs and their distinctive characteristics [1]. The use of large scale systems needs more attention in the area of test and evaluation. During the 1960's enhancements to the subject had come in the forms like a "Pre-processing program for real-time ECG rhythm analysis" [2], real time processing in antenna array [3], frequency analysis in digital computers [4], and so on. Further, the paper explains various techniques for optimum use of the multiprocessing systems in reducing the time for computations [5].

To organize and guarantee real-time performance in a timely manner, efficient scheduling policy is required which can prevent overloading of tasks. Scheduling is the process of allocating resources such as processors, memory and time to the tasks for execution in an effective way. As far scheduling methods are concerned there are various scheduling methods introduced in due course. The initial scheduling method introduced was the "First Come First Served (FCFS) approach". A real-time computing system has to deal with the arrival of soft



and hard deadline tasks for execution. The scheduling approach like FCFS cannot feasibly schedule all the hard real-time tasks before their deadline. Multi-level programming and generalized multiprocessing in multi-programming systems were designed well in the 1960's [49, 50]. This initial approach is classified as pre-traditional approach in this paper. The scheduling approaches which were introduced after the preliminary models are classified into conventional/ traditional RTS (1960-1980), Flexible or adaptive RTS (1981-2000), Current Intelligent RTS (2001 to the current date), and future RTS with improved intelligence and processing abilities.

The paper is explained as follows: In section 2 we present the conventional/ traditional real-time systems and their limitations and challenges. Section 3 discusses improved real-time system by explaining the fault tolerance, feedback control, and value-based scheduling approach used to improve the flexibility in real-time system with its challenges. Section 4 demonstrates the approaches involved in current real-time system. Section 5 explains the future RTS with the mapping of catastrophe theory with intelligent real-time system to maintain stability of the system, when catastrophe happens due to environment change. Section 6 concludes the paper.

2. Conventional/traditional Real Time System Scheduling Approach

There are many challenges faced in modern embedded systems which are time-sensitive. The scheduling policies dealing with the uniprocessor and multiprocessor systems are explained, and their limitations are depicted. In the decade of 1960's and early 1970's, the uniprocessor scheduling scheme served the purpose of guaranteeing real-time performance [9] [51]. The main factors to be considered during scheduling are time criticality of tasks, available resources, and scheduling policies. The distribution of workload in processors is done based on the functional parameters of tasks. The functional parameters are pre-emptive nature of tasks, criticality of tasks, and optional executions [40]. A task is said to be pre-emptive if it can be suspended from its execution for some time allowing another job for execution and later resumed from its point of interruption. A task is nonpreemptable if it cannot allow any interruption and has to be executed from the release time to its finish time. The criticality of the task is an additional parameter that indicates the importance of the tasks among other tasks. This parameter that is an integer is mainly used during overload, where only few high critical jobs are scheduled to execute. Few critical real-time tasks must always comply with its deadlines, even under overloaded conditions. The optional tasks are those tasks that are not mandatory jobs to get executed, and it can either complete its execution in a delayed mode or can never get executed.

2.1 Uniprocessor Scheduling Approach

In a uniprocessor system, the scheduling policies determine the task sequence to execute, and the schedulers implement the same as given in Figure 1. In a classical uniprocessor system, there are three common approaches applied to scheduling. They are clock-driven, weighted round-robin, and priority-driven algorithms [40].

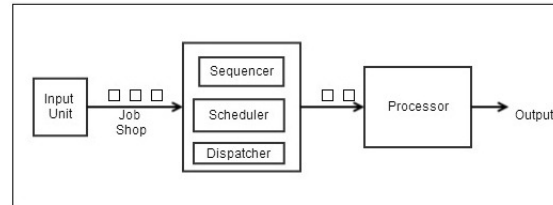


Figure 1. Uniprocessor RTS

Clock-driven scheduling is applicable when decisions are made at specified time instant. The time sharing applications are designed using round-robin approach. Priority-driven algorithms are also called event-driven scheduling, are implemented by assigning priorities to the jobs and it rarely leaves any resource idle. Jobs which are in ready state are placed in a queue ordered by the priorities of the tasks. Deadlines are used to assign priorities to the task. The priority-driven algorithms are fixed priority (FP), Least Laxity First (LLF), Rate monotonic (RM), and Earliest-Deadline-First (EDF) algorithm. In dynamic priority EDF scheduling, the priorities are changed based on their deadline. The theorem stated by Liu in [40] is shown below:

Theorem 1. "When preemption is allowed, and jobs do not contend for resources, the EDF algorithm can produce a feasible schedule of a set J of jobs with arbitrary release times and deadlines on a processor if and only if J has feasible schedules."

For example, assume a system with tasks according to the following table:

Task	Arrival Time	Execution Time	Deadline
T_1	1	9	30
T_2	4	3	10
T_3	5	10	25

At time 1, T_1 arrives at the system. Since it is the first task waiting to run, it starts executing immediately. At time 4, T_2 arrives. When d_2 is compared with d_1 , and since $d_2 < d_1$, T_2 pre-empts T_1 . At time 5, T_3 arrives at the system. When d_3 compared to d_2 and since T_3 has lower priority than T_2 , T_3 waits till T_2 finishes its execution. Then T_3 starts its execution. When T_3 completes its execution, T_1 can be resumed and run to completion.

In [42] the feasibility analysis of EDF states "If the total utilization of the task set is no greater than 1, the task set can be feasibly scheduled on a single processor by the EDF algorithm."



For example, consider a system with the task set as given below:

Task	C_i	T_i	D_i
T_1	1	1	2
T_2	1	2	4
T_3	1	3	8

Here, the tasks are not schedulable even though the utilization is not greater than 1.

$$U = \frac{1}{2} + \frac{1}{4} + \frac{1}{8} = \frac{7}{8} < 1$$

So the utilization based analysis and the response time analysis alone is not sufficient in any real-time systems.

In hard, critical real-time systems the priority driven method is not used because the “timing behaviour is non-deterministic when job parameters vary” [40].

In static priority assignment, the rate monotonic (RM) scheduling algorithm states that a priority is assigned to each task based on its period. Given the computation time, C_i , and period, T_i , for task i , its CPU utilization U_i can be calculated with the following equation (2):

$$U_i = C_i / T_i \quad (2)$$

For rate monotonic scheduling, the processor utilization for n tasks has been shown to be the following equation (3):

$$U(n) = n(2^{1/n} - 1) \quad (3)$$

Liu and Layland in 1973 addressed the scheduling algorithms in a multi-programming hard real-time environment [6]. Usually, the execution of a single algorithm can have only 60 % processor utilization. So, the combination of both the algorithms (EDF & RM) gave higher results. But this analysis and comparison was done on a single processor. Although there are enormous research finding in this area and documented in textbooks, there is still scope for further research [51].

2.2 Multiprocessor Scheduling Approach

Multiprocessor scheduling approach has its origin from early 1970's [12]. The main challenges with multiprocessor scheduling are the allocation problem and the ordering problem. The scheduling policy decides which tasks to be put in the same processor. The survey papers [7, 12, 53, and 54] focus on the scheduling approaches and describe the difficulties of scheduling recurrent tasks with their sufficient feasibility conditions. This scheduling theory starts with the classification of multiprocessor systems. They are mainly classified as a heterogeneous system and homogenous system. The identical processors which can be used interchangeably are called homogeneous processors and the unrelated processors which cannot be interchanged are called heterogeneous processors. The execution speed of each task in different processors may vary in time.

This paper deals with the homogeneous processors to avoid confusion.

The jobs will be sequencing to different processors in a pipeline fashion as given in Figure 2.

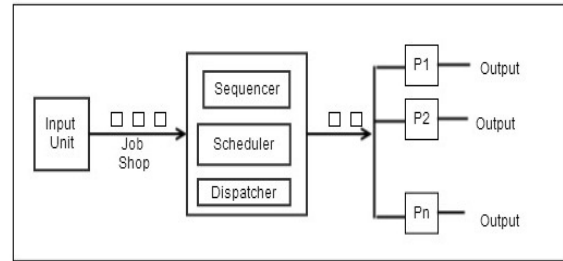


Figure 2. Homogenous Multiprocessor RTS

There are two primary task models: periodic and sporadic task model. The jobs of a task arrive strictly in a periodic manner in periodic task model. The jobs of a task arrive at a random time but after the minimum interval time in a sporadic task model. The task set in the system can be synchronous or asynchronous. The processors (P_1 , P_2 , and P_n) interact among themselves through an interprocessor communication medium via a shared memory. The allocation and schedule in a multiprocessor environment work in the sequence given in Figure 3. The allocation of jobs to processor is done, and the feasibility of the schedule is tested in each processor. If the schedule is feasible, then output the schedule. Otherwise, check the stopping criterion in the system that may lead to either continue making new allocation or declare failure and stop execution.

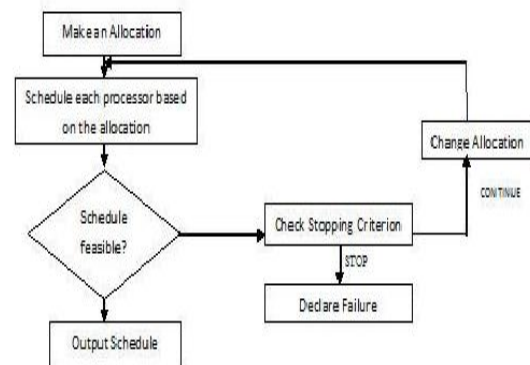


Figure 3. Multiprocessor Schedule

Fault tolerance can be achieved by having multiple versions of tasks on different processors. The task of developing a multiprocessor schedule is, therefore, divided into two steps:

1. Assign tasks to processors
2. Run the EDF(Earliest Deadline First) algorithm to schedule the task allocated to each processor
3. The tasks are assigned to processors based on bin-Packing algorithm where the total processor utilization ($u(i) < \text{threshold}$) must not exceed a given threshold. The constraint behind the EDF is the total processor utilization of all task set should not be greater than 1.



4. So the task assignment Bin-Packing algorithm can be mapped with EDF schedule to make the system fault tolerant with multiprocessor that can be further improved so as to reduce unnecessary pre-emption's including migrations between processors.

The various scheduling approaches in multiprocessor systems are partitioned scheduling, global scheduling, and hybrid scheduling. The comparison of partitioned scheduling with global scheduling and vice versa and its advantages are clearly described by Burns and Davis [12]. Mock and Stankovic stated that the optimal scheduling without a prior knowledge is impossible in a multiprocessor case even with no restriction on pre-emption [7, 9]. Even clairvoyant scheduler cannot feasibly schedule all the jobs in an overloaded situation. So during an overload, some low priority jobs can be discarded by allowing other critical jobs to execute before its deadline. According to a given schedule if the job completes its execution before its given deadline, then the value of a job can be assigned with its execution time. The value is equal to zero if the job fails to complete in time according to the schedule [40]. This is supported by the theorem 2 given by Liu.

Theorem 2. "No on-line scheduling algorithm can achieve a competitive factor greater than 0.25 when the system is overloaded."

The schedulability analysis on hard real time algorithms are covered in few survey reports [9] [10] [11] [12]. In those reports, the performance of homogeneous processors and heterogeneous processors is compared. The outputs of the comparison of executing same task set in different processors are listed as dominance, equivalence, and incomparable [1]. The need for a rescheduling algorithm during unusual arrival of tasks was not highlighted. The acceptance test and schedulability analysis of deterministic offline scheduling, online scheduling, and hybrid scheduling algorithms are presented with their results.

A schedulability analysis is a process of validating whether the application indeed meet all its hard deadliness when scheduled using a specific algorithm. It can be used as an acceptance test at every arbitrary entry of new tasks. The acceptance test may fail with the arrival of significant instant of a task.

The formal specification and proof of important properties of the system and inheritance protocols was presented [17]. In 1988, Stankovic addressed the limitations of the prevailing real-time computing and explored the various other approaches to real-time system [13].

2.3 Challenges

Due to the sudden changes happening in the environment, a system is expected to interact continuously with their environment. Adelantado et.al [16] has defined such a system as a reactive system. Intelligent time critical systems were mainly designed

based on the framework of a system called "SATURNE (Synchronous Any Time Coupling for Real-Time Reasoning Design)." This intelligent system supports reactivity, predictability, and flexibility to timing constraints. The biggest challenge is how adaptability, reactivity and predictability can co-operate for a successful result because strong predictability and flexibility are contradictory requirements. These systems require few additional features like adaptability and reasoning processes to meet the challenging demands. These abilities were limited in conventional real-time systems.

3. Improved Real-Time Scheduling Approach

The tasks are scheduled in offline mode and online mode in conventional real-time system. The system needs to be flexible enough such that they use resources inefficiently and react inflexibly to failures and overloads. Apart from these, the two main responsibilities of a real-time scheduler are

- Perform the acceptance test when new tasks enter the system.
- Dispatch the tasks to processors in a multiprocessor environment.

The output of the test can be of the nature accepted with guarantee, accepted but not guaranteed, and rejected. Based on the output, tasks are scheduled and dispatched to the processors. This decision will be taken at the run time in flexible real-time system. Already many aspects of flexible real-time systems have been addressed before [18, 19, 20, 21, and 22], no systematic approach was given to handle or detect catastrophe. In this paper, we aim to portray and characterize the basis required for detecting and managing catastrophe too.

In a flexible system, a discardable optional function called sieve can be identified by its usage and the decision can be made to take advantage of the possibility of discarding them during overloads.

3.1 Fault tolerant scheduling approach

The motivation for catastrophe management is based on the observation on failures occurred with fault tolerant, flexible real-time systems. Primary-Backup (PB) model is the best-suited model having two copies of a task scheduled on different processors [55]. Fault tolerance can be achieved using redundancy of the task running methods. Here, each flexible job has a primary version and one or more alternate version(s). The primary version produces the accurate result of the desired quality. When the primary version fails, the backup version gets executed. The backups must satisfy the following properties in order to overlap:

- Two different backup copies can overlap in the schedule if their primaries run on different processors as shown in Figure 4 [58].



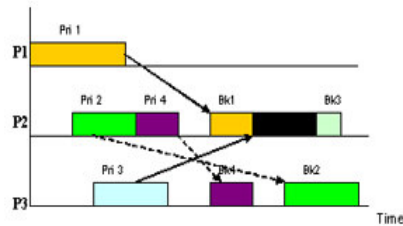


Figure 4. Overlap of backup copies.

Here, the back up copies BK1 and BK3 are overlapped in same processor where as their primary copy are scheduled in different processor.

- The primary copies can also overlap backup copies in the schedule such that there is sufficient slack time for all primary and backup copies of tasks to meet the deadlines.

The quality of the result can be measured in terms of the error in the result or reward attained by the task. The imprecise computation model was used to characterise the flexible system.

3.2 Feedback Control scheduling approach

The dynamic scheduling in real-time system is applicable when the tasks are not known in prior. The online schedule is called dynamic schedule because the unusual arrival pattern of tasks must be scheduled such that all tasks meet its deadline. So the dynamic scheduling algorithms are classified as [19] scheduling with resource sufficient environment and scheduling with resource insufficient environment. As explained earlier, EDF is sufficient for resource sufficient environment. But during overload, EDF's performance degrades. The spring scheduling algorithm was introduced [56] guarantees incoming tasks dynamically through on-line admission control policy. Thus, this is appropriate for resource insufficient environment. The three major paradigm of real-time scheduling during 1960's are RM scheduling, EDF scheduling, and admission control based algorithm. All these three paradigms described earlier are open-loop, which means once the tasks are scheduled then they cannot adapt to any changes based on continuous feedback. So for unpredictable environments, open loop algorithms perform very poorly. The feedback control scheduling (FCS) theory was adapted by real time scheduling [19] to satisfy the required specification of RTS. The FCS scheduling approach has four main components. They are monitor, controller, QoS actuator, and basic scheduler. Few issues with FCS are how a system designer specifies the performance requirement of a flexible real-time system and how a designer systematically designs a scheduling algorithm to satisfy system performance specification [19].

3.3 Value based scheduling approach

The IRIS (Increasing Reward with Increasing Service) model can be used to characterise any time jobs that do not have mandatory component. The reward function associated with an IRIS task increases with the amount of service given to it [42].

Typically the reward function given in [42] is of the form

$$R(x) = \begin{cases} 0 & \text{if } x < m \\ r(x) & \text{if } m \leq x \leq o + m \\ r(o + m) & \text{if } x > o + m \end{cases} \quad (4)$$

where $r(x)$ is monotonically non decreasing in x . The reward is 0 up to sometimes m ; if the task is not executed up to that point, it produces no useful output. Tasks with this reward function can be regarded as having a mandatory and an optional component. The mandatory portion must be allowed to complete its execution before the deadline if the task is critical; the optional portion is done if time permits. The value based scheduling approach guarantees the graceful degradation of the system when failed [23]. Earlier research was focusing on the improvement of job shop scheduling, which has the limitation of taking run-time decisions. The value based scheduling increases the flexibility with guarantee. The causes that can affect the value of service can be based on the environment, computing resource, and the application software. The failures and overload conditions are controlled with the strong admission policy. The admission policy is purely based on the construction and representation of values. Here, the value function will be able to compare two objects using a binary relation [23]. "Let 'More_Useful' be such a relation. Then $S_1 \text{ More_Useful } S_2 \Rightarrow V(S_1) > V(S_2)$ where V is the value function that assigns a real number to the service (it is an approximation of the service utility).

- If V is defined for only a subset of S_i then, it is a partially ordered value function.
- If V is defined for all S_i then, it is an ordinal value function. So,
 $S_1 \text{ More_Useful } S_2 \Leftrightarrow V(S_1) > V(S_2)$
 An ordinal value function is sufficient is sufficient for deciding which single alternative of a service is to be preferred.
- A value function that is ordinal and additive (over the set of services/alternatives) is called a cardinal value function. So,
 $V(S_a \text{ and } S_b) = V(S_a) + V(S_b)$

The challenge in constructing values is to know whether the decision maker's choice is represented correctly using the value function. The best effort scheduling gave by Locke [24] also improves flexibility with no guarantee. Several practical procedures [25, 26] solve the construction problem. The value is derived from AHP (Analytic Hierarchy Process). The first step in AHP is to model the problem in hierarchies. It helps in understanding the problem in all possible criteria, and the alternatives can be defined. Here, the different modes must be listed, and possible criteria can be accounted, and alternative decision task also must be listed to take a decision. The limitation with an autonomous vehicle



control for flexible real-time systems was also addressed [27]. The value was then assigned for each service by having a huge comparison and the decision maker assigns the value. The output of this value based scheduling in an overload situation provides graceful degradation.

The few other algorithms which are used for flexible scheduling are Mandatory-First algorithm, Slack Stealing approach, and so on.

3.4 Challenges

The challenges in adaptive real-time scheduling approaches are

1. Musliner addressed the challenges in real-time AI (Artificial Intelligence) in 1995 [14] [15]. His description of real-time computing is “the system must work fast enough to act on their environments in well-specified ways”. He developed an architecture called CIRCA (Co-operative Intelligent Real-Time Control Architecture) that consists of both Real time and Artificial intelligence co-existing with each other. CIRCA was implemented with the help of many AI algorithms to perform tasks. These algorithms are any-time algorithms. Any-time algorithms mean that they terminate only when they reach the threshold. The speed alone is not sufficient for real-time performance. There are many other aspects of performance [32]. They are responsiveness, timeliness, and graceful adaptation.
2. The system must be flexible without compromising its efficiency and performance. An architecture based on the blackboard paradigm strives to achieve reasoning modules in the system. All the reasoning modules share the blackboard. It provides a way to decouple knowledge processing actions from the data validation actions associated with blackboard data. But the intelligence factor was mentioned as future work.
3. In real-time systems, the involvement of intelligent control is one of the biggest challenges [33]. An intelligent agent can predict the operation based on the environment constraints and its goals. To do this, an intelligent agent must have dynamic entities in the environment. These entities must require knowledge based reasoning skills. So the tasks can occur in various domains such as power plant monitoring and aircraft control, etc.

4. Current Real-Time Scheduling Approach

The flexibility in real-time systems allows taking decisions in the runtime. With the increasing demand in large scale systems, the system needs to be more intelligent to adapt to the changes happening in the environment. Higgins combined “the ability of humans to recognize patterns in data and to make inferences with computer methods for decision-making” [28]. The unusual entry of tasks was

addressed as unplanned events occurring during the scheduled period. The unplanned events are described as different environment modes as DFF mode (day-time, fault free) and PWF (poor, wet and faulty) [23]. When dealing with unplanned events, the scheduling procedures are expected to be ad hoc. In reality, it is impractical to alter the currently scheduled tasks with the newly arrived unplanned tasks. The reason is the scheduled tasks must have already blocked the sufficient resources for its execution. The possible way to maintain stability is by having multiple resources. In large-scale complex system, realizing a copy of resources, usually, will be very costly. The system needs to handle all the circumstances that could ever occur in order to make the system fully automated. Higgins described the hybrid system as the integrated approach which consists of “a knowledge base, an expert system, schedule generator, domain manager, schedule modification system and a human scheduler” [28]. The hybrid scheduler works with the concept of Artificial Intelligence. The human schedulers improve the schedule continuously by providing feedback to the schedule modification system about the current situation. The schedule modification system then learns from the changes and the routine continues. The feedback control scheduling mechanisms are very well addressed in research papers [30, 31].

Many advanced systems provide the graphic representation of the entire working of the system to the scheduler. The human scheduler plays a central role in the decision-making process in hybrid intelligent scheduling as given by Higgins [28]. The complete reassignment of tasks depends on the responsibility and functionality of the person.

Adaptive real-time intelligent systems and its design issues are addressed earlier [29]. Here, real-time environments are controlled and monitored by real time intelligent system. The critical decisions taken by these systems guarantee the safety and improvement of the environment in which they are embedded.

4.1. Mixed Criticality Scheduling Approach

The conventional real-time system provides service only for urgent tasks, whereas the advanced system must also prioritize the high-criticality jobs. Many papers limit with dual criticality levels such as high(HI) and low(LO) [59]. The system is expected to be in either high criticality mode or low criticality mode. Various mode changes were addressed. In each mode few tasks will be scheduled. The mode change protocol addresses that the schedule will be acceptable within the mode and no schedule change is done during the mode change. Slack scheduling, EDF scheduling, shared resources are the few approaches used in response time analysis. In polynomial time, to list the priorities of the mixed criticality jobs Baurah proposed the first efficient approximation algorithm called Own-Criticality Based Priority (OCBP) [41]. The performance of this



algorithm was measured in terms of speed-up factors. Among the on-line scheduling policies, OCBP is proved to have the minimal speedup factor for two criticality levels. In the same way, the mixed criticality sporadic systems have a speedup factor of $4/3$ for the extended version of EDF called EDF-VD (Earliest Deadline First-Virtual Deadline) for its two criticality levels. To reflect its importance over low-criticality tasks, EDF-VD assigns the earlier virtual deadlines to the high-criticality tasks.

In [34, 35], we have designed a super scheduler which can reassign the tasks to different processors when there is an unusual arrival of catastrophic tasks. The stability of the system is also well maintained. The multi-core technique allows the execution of mixed criticality tasks with hard real-time deadlines and not hard deadline tasks [43]. In high integrity applications the development of reliable real-time systems uses diverse replications channels to execute the same set of tasks with no dependency between them [44]. Among the two channels, one is fully functional and the other is partially functional. The fully functional channel will have high reliability and the other will use some static analysis to maintain its position. Failure happens when both channels miss a deadline. The software architecture for both the channels is different. Safety critical systems are required to satisfy the certification requirements from various certification authorities [45].

4.2. Worst Case Execution Time Analysis

The worst case execution time (WCET) is the maximum time limit allowed for the execution of a task. It is represented as a function of a high criticality task or low criticality task. Rather than having a single number as WCET, it can be expressed as a probability distribution (pWCET) [59].

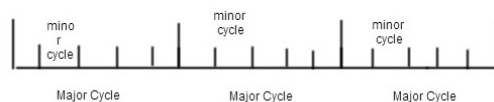


Figure 5. Execution window

The major cycle contains the integral number of minor cycles as shown in Figure 5 for the execution of tasks. The WCET calculated by system designer and the certification authorities will be different. These mixed criticality systems have some scheduling problems which cannot be solved by conventional real-time systems. A mixed criticality job has four tuple parameters [45]. Apart from release time and deadline there are few other parameters to be calculated like criticality of the job and the value of WCET.

The huge gap between the WCET and the actual execution time during the schedulability analysis can be accounted and reclaimed later for less critical work as a “slack time” [45]. This allows redistributing the slack across criticality levels to provide temporal isolation among tasks.

The future problems dealing with timing analysis [57] are increased support for flow analysis, verification of abstract flow models, integration with scheduling, and interaction with energy awareness.

5. Future Real-time Scheduling Approaches

The Probability law is used to analyze the present scenario and to make the future prediction [36]. With the entry of unusual arrival patterns of tasks, the overload condition of the system can be detected by the deviation theory. To detect such scenario, some methods are required to describe the characteristics of catastrophe and its dynamical structure. Catastrophe theory that was introduced by Thom [37] in late 1960's can be used to describe and detect the structure of catastrophe. Thom's idea was developed in various domains. His research indicated that there are seven catastrophic forms in the situation of the control variable less than four. The seven kinds of catastrophe models are “the fold, the cusp, the swallowtail, the butterfly, the hyperbolic point, the ellipse singular point, and the parabola singular point”.

In complex systems, the catastrophe theory is mainly useful in solving the problems. This theory is best suited for all smooth parametric models. So, it is not applicable for any unsmooth system. The complex features of real time system are controlled by multiple factors. In order to maintain the stability of a real-time system, the catastrophe theory can be applied to real-time system scheduling. To do so, the scenario of any real time system can be taken as state variables. The states changes between normal, transient and anomaly. These state changes include changes in reprioritizing the tasks, allocating resources, and executing the tasks with strict timing constraints. This process is called the catastrophe handling. To handle this process, the analysis and identification on the forms of catastrophe theory is mandatory.

In [36], swallow tail model describes the difference set equation and various other control variables affecting the catastrophe. The catastrophe theory was used to detect the network traffic anomaly by adopting the catastrophe series as the fundamental formula. As an outcome, two parameters are identified to evaluate the performance of the detection method. They are “probability of detection and the false alarm rate for the anomaly event”. The ratio between the number of detected anomaly events and the total number of occurring anomaly events indicates the probability of detection. The false alarm detection capability rate is measured by the division of number of false detection points and the total number of all involved detection points. The catastrophe theory can also be used to distinguish different kinds of network traffic anomaly.

In [38], an approach to analyse the effect of catastrophe caused by the unusual entry of tasks from the environment change, was introduced to detect the catastrophe and measure their severity. Off late



catastrophe theories have inclined to concentrate on soft computing. It can also be extended to large scale systems with a minimum of differential calculus. The applications deal with the structure of stability with sudden shifts. Morse's Lemma makes the description of classification of critical points [37]. Critical point is any value in the domain where either the function is not differentiable or its derivative is 0. Morse's Lemma is used to describe the minimum, maximum and the points of inflexion values. Thom's Catastrophe theory drift method and the Potential Functions [39] are applied to workload balancing in a Distributed Database system. Morse critical points have an important stability property which is stated as [37] "preservation of type under small perturbations".

6. Conclusion

The advances in scheduling approaches to real-time systems were addressed from its inception to the current period. The challenges in handling the unusual entry of tasks that may happen in real-time systems were dealt with varying degree of complexity ranging from conventional approaches to intelligent scheduling approaches. The need for an intelligent module to detect the catastrophic occurrence and measure the severity of it in order to improve the flexibility of dealing with it is also highlighted. Later the integration of catastrophe theory with real-time systems implemented through scheduling systems is presented. The seven geometries of catastrophe theory in detecting the eminence of catastrophe and the techniques of maintaining the stability of the system with moorse lemma are presented.

The research in real-time scheduling systems is still active and vigorous though there are several advance people have made to compensate the existing problems in the domain. The challenges in the current field are improving the schedulability analysis and worst case computation complexity analysis, and as a future challenge, the causes for the critical points severity can be measured using the catastrophe theory.

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Biographies



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Implementation of FPGA on SPM Drive Using PI Controller

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Abstract

This paper presents the FPGA implementation of a Direct Torque Control (DTC) for Surface Mounted Permanent Magnet Synchronous Motor (SPM) using PI controller. Reference stator flux is obtained with Maximum Torque per Ampere Principle (MTPA), which gives minimum stator ohmic losses resulting in maximum efficiency. The Direct Torque Control algorithm with PI controller is designed and implemented using VHDL. The complete digital controller algorithm is divided into two modules. From the first module position of flux vector is found based on the flux error, torque error and the sector. In this module the reference torque is obtained using PI controller from which the online reference stator flux is obtained. From the second module the switching state of the inverter is obtained based on the position of the flux vector. The digital controller algorithm is implemented on a Xilinx Spartan-3 FPGA board. The inverter keeps the same state till the outputs of the hysteresis controllers change states. So that when inverter is fed to the SPM it maintains a desired constant speed under different load conditions. Experimental and simulation results for transient response of the drive are tested and compared.

Keywords: Field Programmable Gate Array (FPGA), Surface Mounted Permanent Synchronous Motor (SPM), Maximum Torque Per Ampere (MTPA), Direct Torque Control (DTC), PI controller

Nomenclature

v_β	two axis stator voltage along direct axis, V
v_α	two axis stator voltage along direct axis, V
v_d	direct axis stator voltage, V
v_q	quadrature axis stator voltage, V
i_d	direct axis stator current, A
i_q	quadrature axis stator current, A
R_s	stator armature resistance, Ω
L_d	direct axis inductances, H
L_q	quadrature axis inductances, H
ω_r	rotor speed in electrical rad/s
T_e	electromagnetic torque, Nm
P	no. of poles
λ_f	magnetic flux linkage, wb

λ_s	stator flux linkage, wb
λ_d	direct axis flux linkage, wb
λ_α	two axis flux linkage along direct axis, wb
λ_β	two axis flux linkage along quadrature axis, wb
λ_q	quadrature axis flux linkage, wb
T_L	load torque in Nm
J	moment of inertia in Kg.m ²
Θ	angle between stator and rotor flux linkage
λ_{ref}	reference stator flux, wb
FPGA	Field Programmable Gate Array
SPM	Surface Mounted Permanent Magnet Synchronous Motor
MTPA	Maximum Torque Per Ampere

1. Introduction

The permanent magnet synchronous machine (PMSM) is a synchronous machine where in the excitation winding in rotor is replaced with permanent magnets. Thus resulting in negligible rotor losses and hence an improved efficiency, outstanding power to weight ratio, offer an improved power factor relatively independent of the pole number and speed. The surface mounted permanent-magnet synchronous motor (SPM) also known as the axial flux permanent magnet motor, where, the permanent magnets are placed on the surface of a cylindrical iron-laminated rotor body [1]. The absence of rotor winding and its related losses lead to high efficiency, high torque/weight ratio, and reduced cooling requirements [2].

Also due to high equivalent magnetic air gap the armature reaction effect on pole flux of SPM is low when compared with other machines of similar size [3],[4]. Features like high efficiency, high power density and linear torque characteristics of this motor is made suitable for a wide range of applications like in high performance elevator drive systems, actuators for industrial robots and wheel in motor for hybrid vehicles. The flux-weakening operation with sufficient torque capability of SPM finds applications in wind generators in attaining a wide range of speed control [1].

The control methods of AC drives depend on advanced microprocessor and DSP techniques to implement the complex, real-time control algorithms necessary for high dynamic performance. These conventional techniques has the disadvantages like complexity in design, more power consumption, limited computational capability and hence



a hindrance for rapid prototyping and design of modern electronic products.

The advent of FPGA (field-programmable gate array) technology has enabled rapid prototyping of digital systems. FPGA systems allow easy implementation of digital signal processing due to its higher performances, enhanced flexibility and scalability, lower cost and parallel processing feature leads to less computational time. Many control functions tend to migrate from microcontrollers (or DSP) platforms to SoPCs. [5], [15].

The technology of self-regulating hardware description languages like VHDL and Verilog are used to program FPGAs. The digital FPGA-based solutions have been effectively working for motor control applications [6]. VHDL is a rich and versatile language that can be used for synthesis, modelling, simulation and is supported by all major Computer Aided Engineering platforms and tools which can compile VHDL designs into a large variety of target technologies [7]. In this work, VHDL is used to implement the Direct Torque algorithm for SPM drive.

The FPGA realization of the PWM strategies provides advantages such as fast prototyping, simple hardware and software design, higher switching frequency, reuse, restructure and release the computational load of the microprocessor. Space vector modulation (SVM) is based on the representation of three phase voltages as space vectors. It exhibits the features of good dc-bus voltage utilization and a low THD when compared to other PWM methods [8]. In the proposed digital DTC with FPGA implementation has the following features: very fast dynamic response, less failure chances, reprogrammability, high accuracy (about 99%), high speed [9].

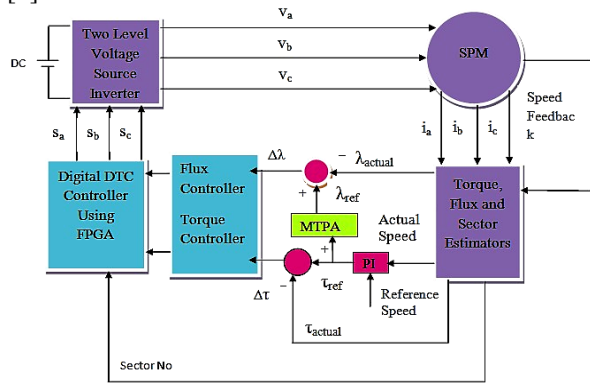


Figure1. Block diagram of the Proposed DTC for SPM Drive

The goal of the proposed DTC is to control the torque which in turn controls the speed of a SPM. Figure 1 represents the generalized components of a SPM drive. The function of the digital DTC controller using VHDL realizes the optimal switching logic to select the appropriate stator voltage vector that will satisfy both the torque status output and the flux status output to obtain the desired speed. The general block diagram of FPGA implementation of a SPM drive is shown in Figure 1.

The organization of the paper is as follows: section 2 describes the modeling of SPM. Section 3 introduces the proposed DTCs with on line reference stator flux

generation using the application of MTPA. Section 4 deals with the dynamic performance of DTC for SPM drive with the application of PI speed regulator to find reference torque and reference flux. Section 5 explains the FPGA implementation of proposed DTC algorithm for SPM drive using PI regulator. Section 6 demonstrates the experimental results. Section 7 concludes the outcome of the paper.

2. Analytical Model of SPM

Mathematical modeling (represented from Equations (1) to (15)) of the SPM is required for simulation of the system and to design a suitable controller. SPM is usually modeled in the rotor synchronous rotating frame (q/d frame), as this approach eliminates the time-varying inductances from the voltage equations in the stator frame. In the rotor reference frame, [10, 11] the voltage equation and the torque equation of SPM are given by equations (1) to (16).

$$v_\beta = \frac{2(v_a - v_b - v_c)}{3} \quad (1)$$

$$v_\alpha = \frac{v_b - v_c}{\sqrt{3}} \quad (2)$$

$$v_d = v_\alpha \cos \theta + v_\beta \sin \theta \quad (3)$$

$$v_q = v_\beta \cos \theta - v_\alpha \sin \theta \quad (4)$$

$$V_d = R_s i_d + L_d \frac{di_d}{dt} - (\omega_r) L_q i_q \quad (5)$$

$$V_q = R_s i_q + L_q \frac{di_q}{dt} + (\omega_r) L_d i_d + (\omega_r) \lambda_f \quad (6)$$

Assuming a linear magnetic circuit, the d and q -axis flux linkages are represented as

$$\lambda_d = L_d i_d + \lambda_f \quad (7)$$

$$\lambda_q = L_q i_q \quad (8)$$

$$\lambda_\alpha = \lambda_d \cos \theta - \lambda_q \sin \theta \quad (9)$$

$$\lambda_\beta = \lambda_d \sin \theta + \lambda_q \cos \theta \quad (10)$$

$$\lambda_s = \sqrt{\lambda_\alpha^2 + \lambda_\beta^2} \quad (11)$$

$$\theta = \tan^{-1} \left[\frac{\lambda_\beta}{\lambda_\alpha} \right] \quad (12)$$

The electromagnetic torque is an important output variable that determines mechanical dynamics of the machine as the rotor position and speed varies, which is represented by the following equation

$$T_e = \frac{3}{2} \frac{P}{2} [L_d i_q i_d + \lambda_f i_q - L_q i_q i_d] \quad (13)$$

As indicated in [1], stable torque control can be achieved if

$$\lambda_s \leq \frac{1}{2} \lambda_f \quad (14)$$

The amplitude of stator flux linkage should be chosen according to the equation (6) if fast dynamic response is



desired. Equation (5) consists of the excitation torque which is produced by permanent magnet flux and the reluctance torque. In SPM the d-axis and q-axis inductances namely L_d and L_q are equal to synchronous inductance (L_s) without saliency, therefore the reluctance torque becomes zero and torque is simplified as

$$T_e = \frac{3}{2} \frac{P}{2} [\lambda_f i_q] \quad (15)$$

The rotor speed, ω_r is obtained from equation (16)

$$T_e - T_L = \frac{2}{p} J \frac{d\omega_r}{dt} \quad (16)$$

3. DTC With On Line Reference Stator

Flux Generation

Direct Torque Control describes the way in which the control of torque and speed which are directly based on the electromagnetic state of the motor, similar to DC motor. DTC is the first technology to control the “real” motor control variables of torque and flux. The function of the DTC is to realize the optimal switching logic to select the appropriate stator voltage vector that will satisfy both the torque status output and the flux status output.

In the proposed DTC, first the reference torque is evaluated using the PI regulator, from which the reference flux is obtained. The reference flux and reference torque are very much essential to select a optimum switching strategy. The inputs to the DTC are flux error, torque error and the position of the flux vector (sector number). To achieve the desired speed control, two hysteresis control loops are taken in DTC. One for control of torque and the other for control of flux. The flux controller controls the machine operating flux to maintain the magnitude of the operating flux at the rated value till the rated speed. Torque control loop maintains the torque to the torque demand. The outputs of these controllers are depending on the instantaneous position of flux vector. The online stator flux reference calculation, flux controller and torque controllers are described below.

3.1. Stator Reference Generation

The reference torque and flux of SPM are calculated based on the torque ripple, flux ripple and desired speed requirement. Further on the torque error and flux errors are calculated to choose appropriate switching voltage vector. In the Proposed DTC (both for case I and case II), “Maximum Torque per unit Current/Ampere (MTPA)” [12] principle is used to find the reference stator flux which in turn is obtained from the reference torque up to rated speed of the drive. This ensures minimum stator ohmic losses resulting in maximum efficiency. For Above Rated Speed of the drive “Maximum Torque per unit Flux (MTPF)” principle is used to find the reference flux.

With MTPA, it is essential to implement a optimal switching strategy which ensures minimum stator ohmic losses and resulting in maximum efficiency. Else the desired electromagnetic torque should be obtained with minimum stator current. Thus i_d is forced to zero by the appropriate switching vectors. The reference flux is obtained as follows:

$$\lambda = \sqrt{(L_d i_d + \lambda_f)^2 + (L_q i_q)^2} \quad (17)$$

With $i_d=0$

$$\lambda = \sqrt{\lambda_f^2 + L_q^2 i_q^2} \quad (18)$$

then electromagnetic torque is given by

$$T_e = \frac{3}{2} p [\lambda_f i_q] \quad (19)$$

From equation (13), i_q is given by

$$i_q = \frac{2T_e}{(3 p \lambda_f)} \quad (20)$$

Substituting i_q in equation (11) yields

$$\lambda = \sqrt{\lambda_f^2 + L_q^2 \left[\frac{2T_e}{(3 p \lambda_f)} \right]^2} \quad (21)$$

$$\lambda_{ref} = \sqrt{\lambda_f^2 + L_q^2 \left[\frac{2T_{e ref}}{(3 p \lambda_f)} \right]^2} \quad (22)$$

Here the reference flux is calculated in the online by considering the instantaneous reference torque from the equation (22). In this equation, the reference torque is obtained by converting the rotor speed error into instantaneous torque using PI controller [12] which is shown in Figure 1.

3.2. Flux Controller

The reference flux which is calculated from (16) and actual value of the stator flux are processed through the two level hysteresis flux controller to find out the status of the voltage vector. The principle of the flux controller is explained below.

i). If $|\lambda_{s-ref}| - |\lambda_{s-est}| > \frac{\Delta \lambda_s}{2}$, then the output of the

flux controller is ‘1’ i.e. it will command to increase the flux magnitude.

ii). If $|\lambda_{s-ref}| - |\lambda_{s-est}| < -\frac{\Delta \lambda_s}{2}$, then the output of

the flux controller is ‘0’ i.e. it will command to decrease the flux magnitude.

Thus the flux controller maintains the magnitude of λ_s , within the limits of $2 \Delta \lambda_s$ [14]. The variation of the estimated flux is shown in the Figure 2.

3.3. Torque controller

The function of the torque controller is to control the torque around the reference value of torque. The reference value of torque, T_{ref} is decided by the speed controller output when the drive is in the speed control mode and it is decided by the torque reference set command when the drive is in torque control mode. Here the reference torque is obtained from PI regulator as shown in Figure 3. This reference flux is processed through the three level torque controller to find out the status of the voltage vector. The principle of the torque controller is explained below.

i). If $(T_{ref} - T_e) > \Delta T_e$ then $S_T = 1$ i.e. increase / decrease the torque by switching the states which will accelerate/ decelerate the λ_s in counter clockwise /clockwise direction respectively.

ii). If $-\Delta T_e < (T_{ref} - T_e) < \Delta T_e$ then $S_T = 0$ i.e. reduces the torque by switching the zero states.



iii). If $(T_{ref} - T_e) < -\Delta T_e$ then $S_T = -1$ i.e. decrease / increase the torque by switching the states which will decelerate / accelerate the λ_s in counter clockwise / clockwise direction respectively. The principle of the three level torque hysteresis controller is shown in Figure 3.

The fast torque response can be obtained by selection of appropriate voltage phasors, which depend on the instantaneous position of λ_s , the output of the flux controller and the torque controller. The general selection of vectors for DTC is given in table 1. Depending upon S_T , S_λ and S_θ the switching state is determined from the look-up table as given table 2. Two cases of space vectors are considered for the implementation of DTC. The sector divisions for these methods are mentioned in table 3. The voltage space vector selection for the second case is shown in Figure 4.

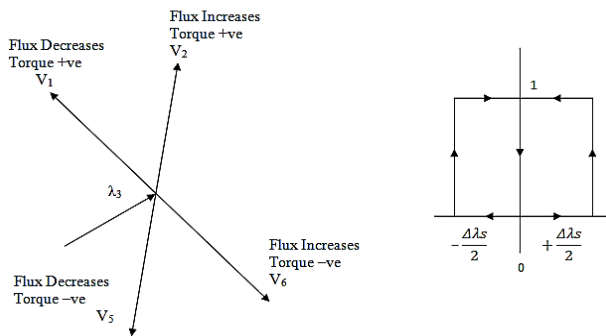


Figure.2 Variation of estimated flux and its hysteresis controller

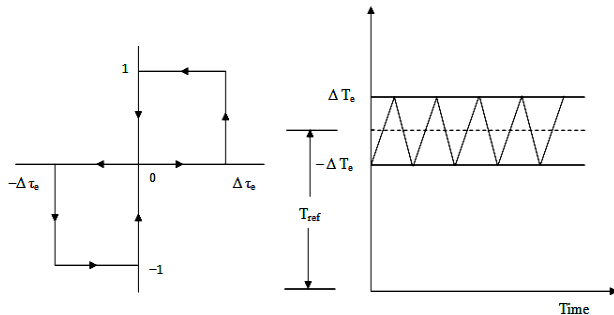


Figure. 3 Three level torque hysteresis controller

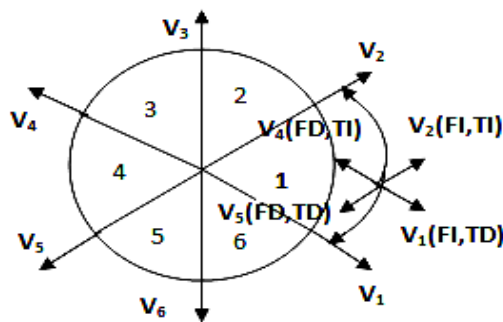


Figure 4. Voltage vector selection for Case II

Table 1 General Selection of vectors for DTC [16]

In the 'K' Sector	Increase	Decrease
Stator Flux	V_{K+1}, V_{K-1}	V_{K+2}, V_{K-2}
Torque	V_{K+1}, V_{K+2}	V_{K-1}, V_{K-2}

Table 2: Selection of Voltage Vector from Torque and Flux Hysteresis Controllers

S_λ	S_T	$\theta(1)$	$\theta(2)$	$\theta(3)$	$\theta(4)$	$\theta(5)$	$\theta(6)$
1	1	V_2	V_3	V_4	V_5	V_6	V_1
1	0	V_8	V_7	V_8	V_7	V_8	V_7
1	-1	V_6	V_1	V_2	V_3	V_4	V_5
0	1	V_3	V_4	V_5	V_6	V_1	V_2
0	0	V_7	V_8	V_7	V_8	V_7	V_8
0	-1	V_5	V_6	V_1	V_2	V_3	V_4

Table 3: Sector selection

Sector No.	Case I	Case II
1	$0^\circ \rightarrow 60^\circ$	$-30^\circ \rightarrow +30^\circ$
2	$60^\circ \rightarrow 120^\circ$	$+30^\circ \rightarrow 90^\circ$
3	$120^\circ \rightarrow 180^\circ$	$90^\circ \rightarrow 150^\circ$
4	$180^\circ \rightarrow 240^\circ$	$150^\circ \rightarrow 210^\circ$
5	$240^\circ \rightarrow 300^\circ$	$210^\circ \rightarrow 270^\circ$
6	$300^\circ \rightarrow 360^\circ$	$270^\circ \rightarrow 330^\circ$

3.4. PI Controller

The dynamic analysis of the SPM when it is transformed to the estimated rotor frame is nonlinear. Under this situation the stability of control system is analyzed as a linearized error model using PI controller. The block diagram of the PI controller is given in Figure 5.

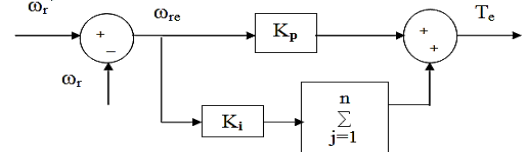


Figure 5. Block diagram of PI speed controller

The output of the controller at n -th instant is as follows:

$$T_{e(n)} = T_{e(n-1)} + K_p \omega_{re(n)} + K_i \omega_{re(n)} \quad (23)$$

Where $T_{e(n)}$ is the torque output of the controller at the n -th instant, K_p is the proportional gain constants and K_i is the integral gain constant. A limit of the torque command is imposed as

$$T_{e(n+1)} = \begin{cases} T_{e \max}, & T_{e(n+1)} \geq T_{e \max} \\ -T_{e \max}, & T_{e(n+1)} \leq -T_{e \max} \end{cases} \quad (24)$$

Here the reference torque is obtained from PI controller under dynamic conditions to achieve desired speed with optimum switching state of the inverter

4. Dynamic Performance of DTC For SPM Using PI Controller

The dynamic performance of DTC for SPM drive is analyzed with two sets of space vectors (Case I and Case II) for different external load torques and for rated speed 3000rpm. The results are tabulated (tables 3 and 4) for Torque Ripple, Flux Ripple and THD Analysis of Phase Currents. When rated load torque 2.5Nm is applied at the rated speed 3000rpm, the torque ripple is 12.21% and



flux ripple is 10.41 % for case I (table3). While for the case II at the same load torque and speed, the torque ripple is 14.81% and flux ripple is 10.63 % (table 4). The current THD is 3.26% for case I and 2.13% for case II for the rated load torque and rated speed respectively. Figures 6 and 7 describe the electromagnetic torque response, speed response and stator flux of SPM drive. From the Fig 6 it is evident that constant speed is obtained under different load conditions. That is 20%, 40%, 60%, 80% and 100% of the rated load torque. Similarly Figure 7 shows the results obtained for constant load torque versus various desired speeds at different instants.

Three phase current wave forms for 3000rpm with 1,5Nm load is represented in Figure 8. The torque response with step change in load from no load to 0.5 Nm for case II is depicted in Figure 9. The load 0.5 Nm is applied at 0.7 sec. The torque response reaches the steady state in 0.7025 sec. This shows that the dynamic response of the drive is faster with the proposed DTC followed by MPTA strategy. The torque ripple and flux ripple are given in Figure 10 and 11 respectively for the same step changes in load. Step changes in speed with constant load are observed in Figure 12.

Table 4: Rated speed = 3000rpm Case I ($0^\circ \rightarrow 60^\circ$)

Time at which load is applied (sec)	Load Torque (Nm)	Torque Ripple (%)	Flux Ripple (%)	RMS Phase Current (A)	Phase Current (THD) %
0.7	0.5	38.46	12.37	1.034	5.10
1.5	1.0	18.18	11.45	1.117	4.79
2.5	1.5	16.46	11.57	1.265	3.71
3.5	2.0	15.59	11.52	1.47	3.63
4.5	2.5	12.21	10.41	1.622	3.26

Table 5: Rated speed = 3000rpm Case II ($-30^\circ \rightarrow +30^\circ$)

Time at which load is applied (sec)	Load Torque (Nm)	Torque Ripple (%)	Flux Ripple (%)	RMS Phase Current (A)	Phase Current (THD) %
0.7	0.5	33.33	12.45	1.085	3.92
1.5	1.0	25	11.52	1.179	3.83
2.5	1.5	22	12.20	1.318	3.66
3.5	2.0	21	12.04	1.511	2.39
4.5	2.5	14.81	10.63	1.767	2.13

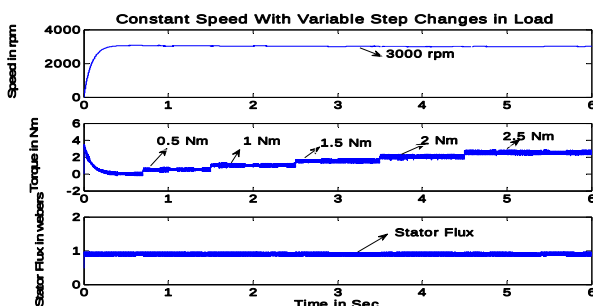


Figure 6: Flux, electromagnetic torque and speed responses for 3000rpm under different loads for Case I

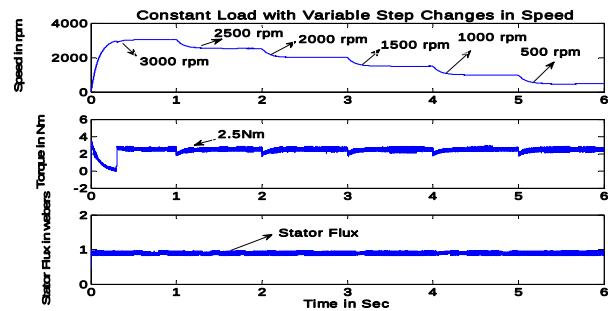


Figure 7: Flux, electromagnetic torque and speed responses for different speeds with constant load for Case I

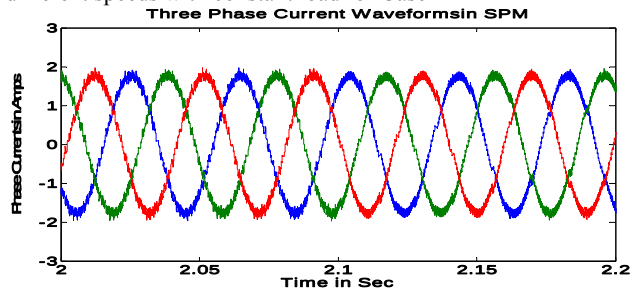


Figure 8: Three phase current waveforms for 3000rpm with load torque 1.5 Nm for Case II

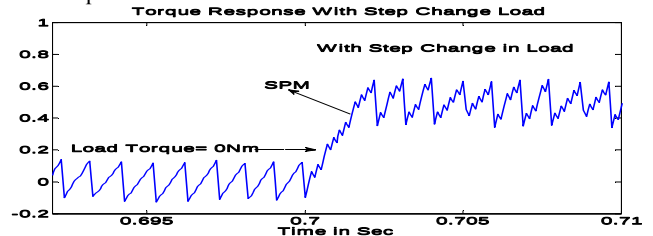


Figure 9: Torque response for 3000rpm with step change in load from 0 Nm to 0,5 Nm for Case II

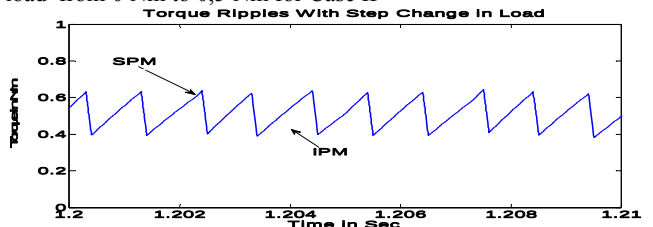


Figure 10: Torque ripple for 3000rpm with step change in load from 0 Nm to 0,5 Nm for Case II

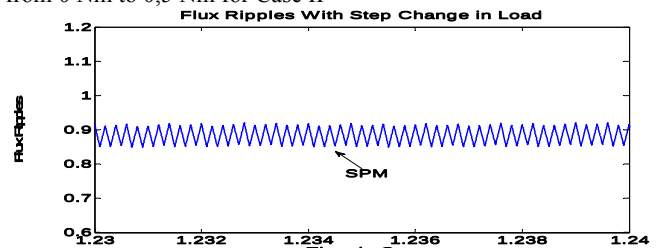


Figure 11: Flux ripple for 3000rpm with step change in load from 0 Nm to 0,5 Nm for Case II

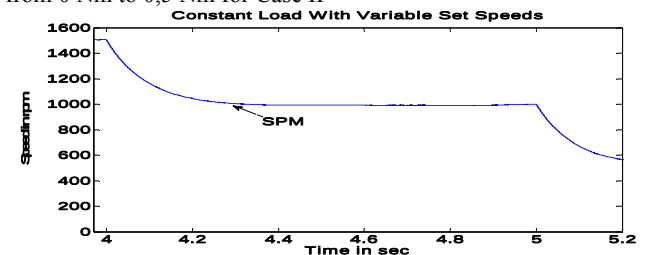


Figure 12: Speed response with variable set speeds at constant load torque for Case II



5. FPGA Implementation of DTC Algorithm

Based on simulation analysis hardware implementation of a SPM using FPGA is carried out for the Case I. The block diagram of FPGA implementation of DTC is shown in Fig 13.

Digital DTC algorithm is realized with Xilinx IST 10.1 simulator and implemented in Xilinx Spartan 3 FPGA board. As already mentioned input to the FPGA board are flux error, torque error and the position of the flux vector (sector number). The flux error represents a two bit binary number (0 or 1 binary equivalent is 00 or 01) which are given to the pin numbers P39 and P40. The torque error represents a two bit binary number (0 or 1 or -1 binary equivalent is 00 or 01 or 11) which are given to the pin numbers P50 and P51. The sector number represents a three bit binary number (0 to 6).

The DTC algorithm with FPGA is divided into two modules. The input to this module are flux error, torque error and sector number. It consists of digital representation of flux controller, torque controller and look up table. From this look up table gives the selection of voltage vector is obtained (table 1). The torque error

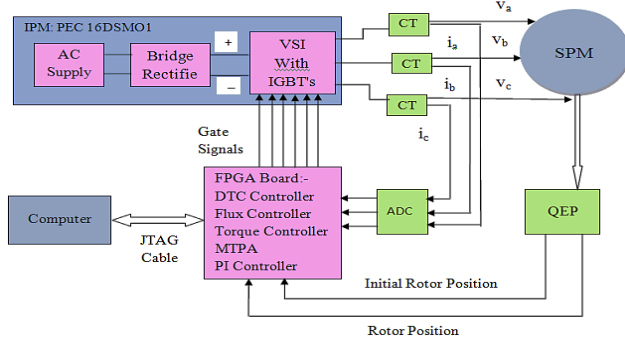


Figure 3. FPGA implementation of proposed DTC for SPM Drive

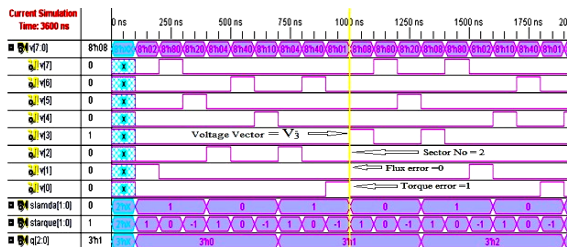


Figure 14 Switching states of the Inverter from sectors 1 to 3 for Module II



Figure.15 Experimental Set up of the SPM Drive with FPGA Implementation

Table 6: Specification of SPM

Parameter Description	Value
Rated Output Power	800 W
Rated Speed	3000 rpm
Rated Voltage	230 V
Torque Constant $\pm 5\%$	0.455Nm/A
Voltage Constant $\pm 5\%$	31.8V/Krpm
Phase Resistance $\pm 5\%$	19.4 Ohms
Phase Inductance $\pm 5\%$	3.82mH
Electric Time Constant	5.6ms
Mechanical Time Constant	0.65ms
Rotor Inertia	1.16kgcm ²

and flux errors are generated when the reference values and actual values of torque and flux are processed through a summer. Here the reference torque is obtained from PI controller, from which the reference stator flux is obtained. From the first module, the position of voltage vector is found based on the flux error and torque error with respect to the sector number.

From Module II, the switching state of the inverter is found based on the position of the flux vector which represents the switching state of the inverter, which is taken from the pin numbers P44, P46 and P48 of Xilinx Spartan 3 FPGA board. The flux error represents a two bit binary number (0 or 1 binary equivalent is 00 or 01). The torque error represents a two bit binary number (0 or 1 or -1 binary equivalent is 00 or 01 or 11). The sector number represents a three bit binary number (0 to 7).

The digital controller algorithm represented here is designed and implemented on a Xilinx Spartan-3 FPGA board with VHDL using Xilinx IST 10.1 simulator. (Device: XC3S400, Pin package : PQ208). In the first step, the complete DTC algorithm is simulated at the transfer level to validate the design. To optimize the design logic, synthesis is carried out in the second step. To generate the FPGA implementation file / bit file, placing and routing is carried out in the third step. Finally the bit file is down loaded in to the FPGA board for testing. Here for example, the results are obtained when flux error is 0, torque error is 1 and the position of the flux vector is in sector 2 (From Table 1) then the switching voltage vector is $V_3(010)$ i.e. the switching states to the inverter are as shown in Fig.14. After this switching state inverter switching state changes to $V_4(011) \rightarrow V_5(001) \rightarrow V_6(101) \rightarrow V_1(100) \rightarrow V_2(110) \rightarrow V_3(010)$ and so on until torque and flux commands change from the existing state.

The complete experimental set up of FPGA implementation of DTC for SPM using PI Controller is shown in Fig.15. This set up consists of 25A/1200V intellectual power module (Model: PEC16DSMO1), 800W surface mounted permanent magnet motor (Model: TC802.821), 5A/5V current transformer, ADC ($\pm 10V$, Bipolar, 12bit, 2MBPS, 8 Channel), 20 MHz clock frequency FPGA board, 6MHz data frequency JTAG cable, Speed encoder (QEP: Quadrature Encoder with Pulses of 2000/500 revolutions per second). The switching frequency of the Inverter IGBT's are 15 KHz with 6 μ sec as the dead time for gate pulses, single phase auto transformer, personal computer and digital oscilloscopes. The technical specification of SPM is represented in table 6.



6. Experimental Results

The motor is fed with a two level three phase voltage source inverter which is implemented by an intellectual power module including gate drivers, six insulated IGBTs and protection circuits. The actual motor phase currents are measured by current sensors (current transformer) which is fed to the control computer through a 12 bit bipolar successive approximation ADC with 1micro sec speed. First only two motor phase currents are measured, as the motor neutral is isolated so only two sensors are required to measure phase currents. The rotor position is measured by means of QEP with 500 pulses per revolutions. To observe the load torque disturbances Spring balance load is coupled to the shaft of SPM.

Experiment is carried out with different sets of load torques and different sets of reference speeds. The electromagnetic torque response, stator flux response and reference speed responses are shown in Fig.16 and Fig17. Further it can be observed from these figures that, for an instantaneous change in the set point of the speed and load torque the tracking performance is very fast and accurate.

The torque ripples are around 18% for SPM drive with the proposed DTC (Case II) using PI controller with FPGA implementation for rated speed 3000rpm and load torque 2.0 Nm. Fig. 18 shows the torque ripple and flux ripple for a rated speed of 3000 rpm and for a load torques of 1.5 Nm and 2.0Nm. From the table 6, the THD of phase currents are 7.199% the THD of line voltage waveforms are 27.997 %; rms phase currents are 2.68A and rms phase voltage 103.8v for the reference speed of 1500 rpm and 3000rpm at a load torque of 2 Nm corresponding values are given in table 7.

From experimental results it is clear that the proposed DTC with PI controller for rated speed 3000rpm and for a load torque 2 Nm gives currents THD 4.5%, where as torque and flux ripples are 18.23% and 20% respectively (table 6). Fig 19. Shows the synthesis result of the proposed DTC using the Xilinx ISE 10.2. The device utilization summary is given in table 8.

The comparative results obtained from simulation for the proposed DTC with online reference flux calculation against torque ripples, flux ripples and phase current THDs are represented in Fig 20, 21 and 22 with case I, II for different load torques and rated speed.

While the hardware results obtained at 60% and 80% of the load torques at rated speed for case I are compared with simulation results (Fig 20, 21 and 22).

From this, it is clear that the torque ripples, flux ripples and THD for phase currents are less for case I. Also it is evident that Two Level Inverter with the proposed DTC, online reference flux calculation, the obtained ripples and THD are less when compared to three level NPCI with online flux calculation proposed in [12].

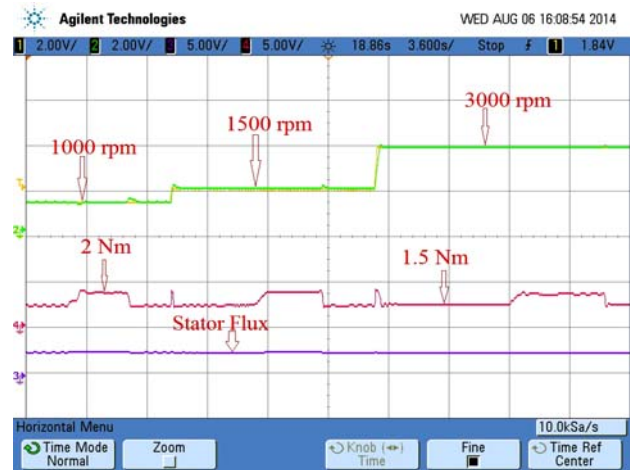


Figure .16 Flux, electromagnetic torque and speed responses for 1000rpm ,1500 rpm and 3000 rpm

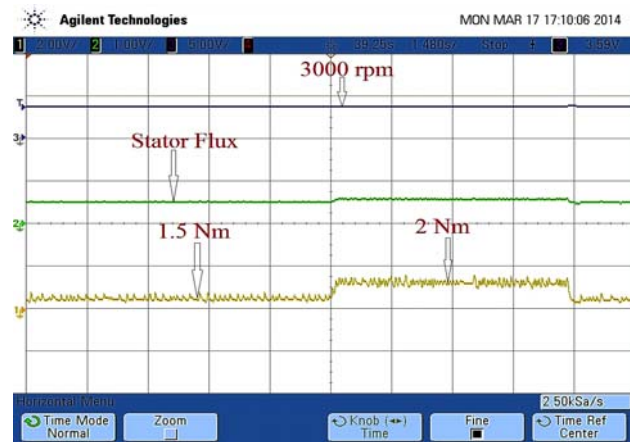


Figure .17 Flux, electromagnetic torque and speed responses for 3000 rpm with load torques 1.5 Nm and 2.0 Nm

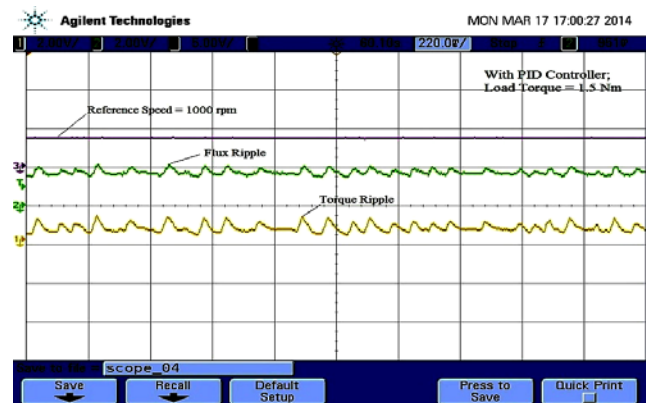


Figure. 18 Torque Ripple, Flux Ripple (reference Speed 1000rpm, load torque 1.5 Nm)

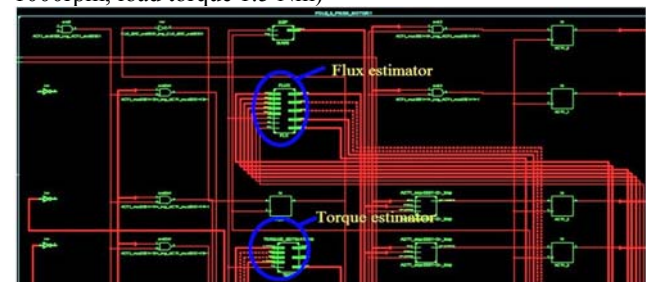


Figure 19. Synthesis result of the proposed DTC using the Xilinx ISE 10.2



Table.7 : Torque Ripple, Flux Ripple, THD Analysis for Line Voltages and Phase Currents of the SPM with FPGA Implementation for Case I using PI Controller

Load Torque (Nm)	Speed (rpm)	RMS Phase Current (A)	RMS Phase Voltage (V)	Phase Voltage (THD) %	Phase Current (THD) %	Torque Ripple (%)	Flux Ripple (%)
1.5	1000	2.68	103.80	27.99	7.19	28.53	20
	1500	2.62	123.99	15.24	3.91	22.34	
	3000	2.70	163	27.76	4.6	20.05	
2.0	1000	5.11	110.02	22.68	4.7	24.87	15
	1500	5.16	130.27	14.94	4.6	22.54	
	3000	5.20	173	28.56	4.02	18.23	

Table 8. Device utilization summary

Device Utilization Summary (Estimated Values)			
Logic Utilization	Used	Available	Utilization
Number of Slices	5415	16640	32%
Number of Slice Flip Flops	3104	33280	9%
Number of 4 input LUTs	8943	33280	26%
Number of IOBs	56	519	10%
Number of GCLKs	3	24	12%
Number of DSP48s	66	84	78%
Xilinx X Power Analysis- Controller Design			
Total Power (in Watts)	0.056		

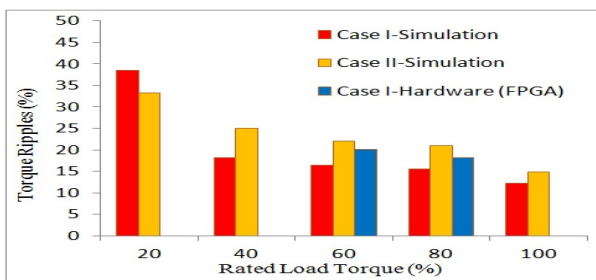


Figure.20. Comparison of the torque ripples for case I and case II.

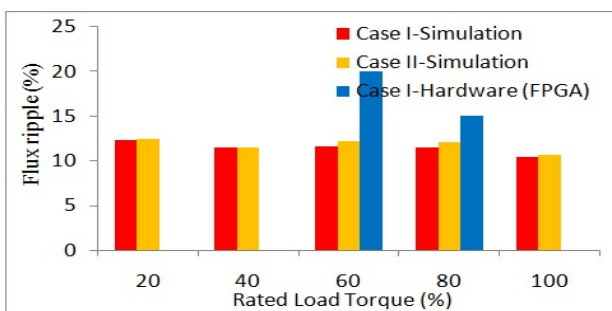


Figure.21. Comparison of the flux ripples for case I and case II.

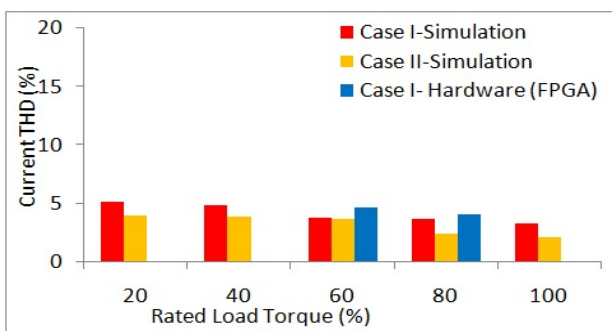


Figure 22. Comparison of the Current THD for case I and case II.

7. Conclusion

In this paper experimental implementation of proposed DTC with online reference stator flux using PI controller for SPM with FPGA by VHDL is carried out. Simulations as well as experimental results are presented to verify the effectiveness of the proposed scheme. The experimental results are compared with simulation analysis and the results are in compliance.

The torque ripple, flux ripples and phase current THDs are minimized with case I for rated speed and different load torques. Since this method gives an effective torque and flux of the drive, it is valid for both dynamic and steady state operations.

With DTC, online reference stator flux the flux hysteresis band can be obtained as minimum as possible which results in reduction of the switching losses by appropriate switching state of the inverter. From the results it is observed that the efficiency of the drive is better even at low speeds.

The scope of the work may be carried out further in reduction of torque ripple, flux ripple, voltage and current THD waveforms by incorporating three level SVM of DTC algorithm using Fuzzy Logic Controller, Genetic Algorithm along with Model reference Adaptive Controller.

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A Computer-aided Optical Pattern Correlation System for Fingerprint Recognition

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Abstract

Pattern recognition is a main topic in the field of biometrics. In such field, pattern recognition needs to be performed at high speed with a high confidence level. Because of the excess and complexity of calculations, traditional digital pattern recognition techniques are not always the best solution especially when real time decision is needed.

In this paper, we propose an optical pattern correlation system for fingerprint recognition. The system utilizes the joint transform correlator (JTC) and the hybrid fringe-adjusted joint transform correlator (FJTC). The input images of fingerprints were correlated directly with the reference fingerprints and a highly distinctive correlation signal was produced if reference and target images are similar, while a negligible correlation signal was produced for different reference and target images. The obtained results showed that optical pattern recognition system is superior in its speed of processing. The accuracy depends greatly on the degree and nature of imposed noise.

Keywords: Optical correlation, joint transform correlator, fringe-adjusted joint transform correlator, fingerprints.

1. Introduction

Nowadays, biometrics is one of the most important field of life and an interesting research. Many biological data are used in biometrics such as DNA sequences and images of different parts. Images include those of the faces, retinas, irises and fingerprints. Fingerprints are the most widely type of images used in biometrics since it has been used more than a century ago. In the past, fingerprints were matched by eyes of experts. Many decades ago, fingerprints has been matched electronically by pattern recognition system. Generally speaking, pattern recognition uses two broad techniques: template matching and classification. In any classification problem, the process is done in two major steps. The first stage is feature extraction and the second stage is classification. The features are

extracted from the classes of a learning dataset should be differ from one class to the other such that the system can use them to differentiate between different classes. Such features are commonly called discriminant features. To determine the class to which a new pattern belongs, same set of discriminant features should be extracted from such pattern [1]. During the classification process, the distances between the features of the new pattern and those of different classes of the learning dataset are measured. The new pattern will belong to the class that has the shortest distance from. Many methods to measure the distance between a pattern and a class are found [2]. These methods differs in their processing time and accuracy and also according to the number of features to be extracted and processed.

In template matching, correlation between an input pattern and other possible patterns is calculated either in the spatial or frequency domains. The patterns that has the highest correlation with the input pattern is the most similar one[3]. Correlation requires a huge number of calculations to be achieved. convolution with its high calculation complexity is the staring step in correlation calculation in the spatial domain. According to the Fourier transform properties, we can correlate to functions in the frequency domain where the convolution process is converted to multiplication process. To correlate to patterns in the frequency domain, we should obtain their Fourier spectrum first. Also, obtaining the Fourier spectrum requires a huge number of calculations.

In case of fingerprints (or any type of two-dimensional images), we apply Fourier transform to tow dimensional data which increases the calculation complexity. Using Fourier transform, the correlation $c(x, y)$ of two-dimensional image $t(x, y)$ and $r(x, y)$ is given by:

$$c(x, y) = FT^{-1}\{T(u, v)R^*(u, v)\} \quad (1)$$

Where FT^{-1} is the inverse Fourier transform, $T(u, v)$ and $R(u, v)$ are the Fourier transforms of $t(x, y)$ and $r(x, y)$ and $R^*(u, v)$ is the complex conjugate



of $R(u, v)$. The Fourier transform of $t(x, y)$ is given by:

$$T(u, v) = \int_{-\infty}^{\infty} \int_{-\infty}^{\infty} T(x, y) e^{-j\pi(ux+vy)} dx dy \quad (2)$$

Where u and v are the spatial frequencies [4].

So, pattern recognition is computationally exhaustive and time consuming process. Knowing that Fourier transform of an image could be obtained at the speed of light using lenses or optical correlation, real time processing come to mind.

In medicine, pattern recognition needs to be performed at high speed with a high confidence level. Because of the excess and complexity of calculations, traditional digital pattern recognition techniques are not always the best solution especially when real time decision is needed.

Because of the parallelism and high speed of optical calculations, optical pattern recognition is considered as the most proper matching method. Optical correlation techniques via the lens Fourier transformation property are the backbone of optical pattern recognition systems. The strength of the optical pattern recognition systems comes from the fact that Fourier transformation, correlation and convolution of two-dimensional signals can be performed in parallel. These properties have always provided optical information processing systems an advantage over electronic systems which are sequential in nature.

In this paper, a description of the optical and hybrid pattern recognition system that we have built in our lab is provided. These include an optical pattern recognition system using JTC and another for FJTC for real-time target recognition applications. The results of trying these systems with a set of fingerprints as a common input of biometrics are promising.

This paper is organized as follows: a review about the work previously done in the area of optical pattern recognition is provided in the following section. Section III presents an explanation of the material and methods used to achieve this work. The results are presented and discussed in the fourth section of this paper. The paper ends with a conclusion and future work.

2. Review

Pattern recognition plays an essential role in the medical field, where it is used to determine and identify the desired pattern or target in an input scene. In digital pattern recognition technique, the input scene must be processed serially, which requires immense computation and ending by having a relatively slow results. On the other hand, Optical techniques have clear advantages in real time application due to its inherent parallel processing. Optical correlator is a device for comparing two

signals through the usage of the lens Fourier transforming properties. As it compares all the pixels in the input scene to all pixels in the reference image at the same time. This parallel processing capability of optical correlators gives high capacity and high speed advantages compared with pure digital processing [5, 6].

Optical correlation can be accomplished using one of two techniques; Vander Lugt Correlator (VLC) [7] and Joint Transform Correlator (JTC) [8]. The advantage of JTC does not need a priori filter fabrication and the strict alignment requirement of VLC. Although the classical JTC has many features compare to VLC, it still has many flaws, e.g., dc (or auto-correlation), sidelobes and large correlation peak width [9]. There have been many attempts to overcome and enhance this flaws or shortcoming. Javidi and Kuo [10] proposed a Binary JTC (BJTC), where the JPS is output binarized to two values (+1 and -1) before applying the inverse Fourier transform operation. BJTC is found to yield higher correlation peak, correlation width and discrimination sensitivity. The main problem of a binary JTC is its relatively long processing time needed for the determination of the threshold value used for the binarization task. Feng and Zhao and Xia [11] proposed an amplitude-modulated filter (AMF) based JTC, where the JPS is multiplied by the AMF before applying the inverse Fourier transform to yield the correlation output, yielding a better correlation performance than the classical BTJ as well as the BJTCs. T Normura [12] proposed a Phase-encoded Joint Transform Correlator (PJTC), where a phase mask is introduced in the input and JPS is multiplied by the same phase mask, this operation preserves the cross correlation component between the reference image and the test image while scattering all other components. The phase-encoding based approach is used to reduce the influence of the extraneous signals such as the zero order term and the conjugate cross-correlations between reference and input targets. In order to sharpen the correlation peaks, Slow operation Phase-encoding of input scene Wide side lobes Sensitive to noise. Alam et al. [13] proposed FJTC where the JPS is multiplied by a real-valued fringe-adjusted filter (FAF) before applying the inverse Fourier transform to yield the correlation function. This technique is found to yield significantly better correlation output than the classical JTC, BJTC and PJTC for input scenes involving single as well as multiple objects. It also avoids the computation-intensive Fourier-plane joint power spectrum binarization processing of a binary JTC.

In this paper, we build an optical pattern recognition system using JTC and FJTC for real-time target recognition applications. this system is used for medical pattern recognition and biometrics and compare between them. The JTC based system starts by displaying both the target and the reference images side by side on an electrically addressed spatial light



modulator (EASLM) placed in a front focal plane of a Fourier-Transforming (FT) lens. By brightening the EASLM using laser, the displayed joint images are optically Fourier-transformed by the lens. Though, a joint power spectrum (JPS) is produced at the back focal plane of the lens, which is then recorded using a CCD camera. the correlation of the input images formed using optical Fourier transformation can be then displayed on an EASLM. For fringe-adjusted joint transform correlator (FJTC), the JPS is then multiplied FAF before inverse Fourier transform operation. We used the images of fingerprints to prove the feasibility of such systems in matching medical patterns. The input images of fingerprints were directly correlated with the reference fingerprints. Though, a high distinctive correlation signal is produced if reference and target image are similar, while a negligible correlation signal is produced in case of different reference and target images.

3. Theory and Methods

Theory of Joint Transform Correlator

The theory of optical correlator is provide in [14] and explained here using figure1. As could be noticed from figure1, $r(x_1 + a, y_1 + b)$ represent the reference image function and $t(x_1 - a, y_1 - b)$ represent the target-image function. Considering the spatial light modulator (SLM) [15]. In this analysis, the electronically-addressed SLM (EASLM) is placed at the front focal plane of lens in the input plane $P_1(x_1, y_1)$ and assuming that the reference image and the input target image are placed side by side on the EASLM. The joint input image function $f(x_1, y_1)$ can be mathematically expressed as:

$$f(x_1, y_1) = r(x_1 + a, y_1 + b) + t(x_1 - a, y_1 - b) \quad (3)$$

Where a, b are the positions of both the reference and target images at the input $P_1(x_1, y_1)$. The distances between the two images must be large enough (greater than the largest dimension in the images) to avoid the overlap between the zero-order peaks and the cross-correlation sidelobes peaks [14].

By illuminating the EASLM with laser, the lens performs the Fourier transform of the input joint image of equation (1) to yield:

$$F(x_2, y_2) = \frac{1}{j\lambda f} T\left(\frac{x_2}{\lambda f}, \frac{y_2}{\lambda f}\right) \exp\left(j2\pi\left(\frac{a}{\lambda f}x_2 + \frac{b}{\lambda f}y_2\right)\right) + \frac{1}{j\lambda f} R\left(\frac{x_2}{\lambda f}, \frac{y_2}{\lambda f}\right) \exp\left(-j2\pi\left(\frac{a}{\lambda f}x_2 + \frac{b}{\lambda f}y_2\right)\right) \quad (4)$$

Where $T\left(\frac{x_2}{\lambda f}, \frac{y_2}{\lambda f}\right)$ and $R\left(\frac{x_2}{\lambda f}, \frac{y_2}{\lambda f}\right)$ correspond to the Fourier spectrum of the target and the reference respectively, λ is the wave length of collimating light, f is the focal length of lenses. The intensity of the complex light distribution produced at the back focal plane of the lens (at the Fourier frequency plane $P_2(x_1, y_1)$), is called the joint power spectrum (JPS), where the joint transform power spectrum (JTSPS) is detected by a square-law detector such as a CCD camera. The JPS is given by:

$$I(u, v) = |F(x_2, y_2)|^2 = F(x_2, y_2) \cdot F^*(x_2, y_2) = \frac{1}{j\lambda^2 f^2} \left[\left| T\left(\frac{x_2}{\lambda f}, \frac{y_2}{\lambda f}\right) \right|^2 + \left| R\left(\frac{x_2}{\lambda f}, \frac{y_2}{\lambda f}\right) \right|^2 + R\left(\frac{x_2}{\lambda f}, \frac{y_2}{\lambda f}\right) T^*\left(\frac{x_2}{\lambda f}, \frac{y_2}{\lambda f}\right) e^{-j4\pi\left(\frac{a}{\lambda f}x_2 + \frac{b}{\lambda f}y_2\right)} + R^*\left(\frac{x_2}{\lambda f}, \frac{y_2}{\lambda f}\right) T\left(\frac{x_2}{\lambda f}, \frac{y_2}{\lambda f}\right) e^{+j4\pi\left(\frac{a}{\lambda f}x_2 + \frac{b}{\lambda f}y_2\right)} \right] \quad (5)$$

Where F^* , T^* and R^* are the complex conjugates of F , T and R respectively.

By expressing the complex field distribution $R\left(\frac{x_2}{\lambda f}, \frac{y_2}{\lambda f}\right)$ and $T\left(\frac{x_2}{\lambda f}, \frac{y_2}{\lambda f}\right)$ into their amplitude and phase distributions

$$R\left(\frac{x_2}{\lambda f}, \frac{y_2}{\lambda f}\right) = \left| R\left(\frac{x_2}{\lambda f}, \frac{y_2}{\lambda f}\right) \right| \exp\left\{j\phi_R\left(\frac{x_2}{\lambda f}, \frac{y_2}{\lambda f}\right)\right\}$$

And

$$T\left(\frac{x_2}{\lambda f}, \frac{y_2}{\lambda f}\right) = \left| T\left(\frac{x_2}{\lambda f}, \frac{y_2}{\lambda f}\right) \right| \exp\left\{j\phi_T\left(\frac{x_2}{\lambda f}, \frac{y_2}{\lambda f}\right)\right\}$$

The joint power spectrum can be rewritten as

$$I(x_2, y_2) = \left| T\left(\frac{x_2}{\lambda f}, \frac{y_2}{\lambda f}\right) \right|^2 + \left| R\left(\frac{x_2}{\lambda f}, \frac{y_2}{\lambda f}\right) \right|^2 + 2R\left(\frac{x_2}{\lambda f}, \frac{y_2}{\lambda f}\right) T\left(\frac{x_2}{\lambda f}, \frac{y_2}{\lambda f}\right) \times \cos\left\{4\pi\left(\frac{a}{\lambda f}x_2 + \frac{b}{\lambda f}y_2\right) + \phi_R\left(\frac{x_2}{\lambda f}, \frac{y_2}{\lambda f}\right) - \phi_T\left(\frac{x_2}{\lambda f}, \frac{y_2}{\lambda f}\right)\right\} \quad (6)$$

The JPS redisplaying on SLM and the lens as shown in Figure1 performs an inverse Fourier Transform of this JPS to yield the correlation output at plane $P_2(x_1, y_1)$, which is detected by CCD camera:

$$c(x_3, y_3) = r(x_3, y_3) \otimes r(x_3, y_3) + t(x_3, y_3) \otimes t(x_3, y_3) + t(x_3, y_3) \otimes r(x_3, y_3) \delta(x_3 - 2a, y_3 - 2b) \quad (7)$$

where $\otimes$ represents correlation operation. The first and second terms are representing the DC auto-correlation term of the reference and target image, as it appears on axis of the correlation plane (x_3, y_3) . The third and fourth terms of equation (5) denotes the cross-correlation terms centered at the position $(x_3 = 2a + 2, y_3 = -2a - 2b)$.



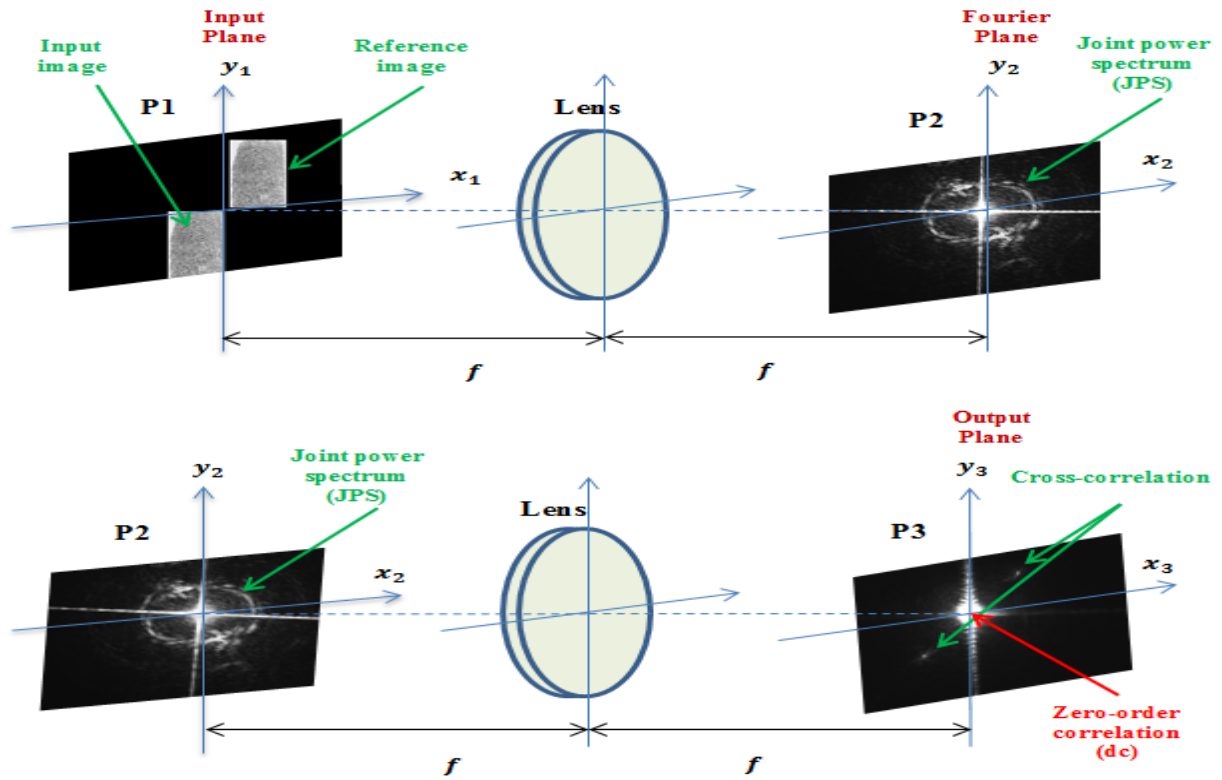


Figure1. Scheme showing the JTC optical architecture used in our simulations. In the upper sketch recording of the joint power spectrum, while the lower sketch shows the obtaining the correlation output.

Fringe-Adjusted JTC

In this technique, the JPS is multiplied by a real valued filter referred to as Fringe Adjusted Filter (FAF) before applying an inverse Fourier Transform to yield the correlation output[16]. The FAF can be given by:

$$H_{FAF(u,v)} = \frac{B}{A + \left| R\left(\frac{x_2}{\lambda f}, \frac{y_2}{\lambda f}\right) \right|^2} \quad (8)$$

Where A and B are parameters taking several forms, where they are used to overcome the pole and suppress noise and control the optical gain respectively. By combining both the JPS and FAF filter, the modified JPS is obtained.

When $B = 1$ and $\left| R\left(\frac{x_2}{\lambda f}, \frac{y_2}{\lambda f}\right) \right|^2 \gg A$ then equation (6) can be simplified as

$$H_{FAF(u,v)} \approx \left| R\left(\frac{x_2}{\lambda f}, \frac{y_2}{\lambda f}\right) \right|^2 \quad (9)$$

By multiplying the JPS with the FAF then the FJPS given as:

$$\begin{aligned} \text{FJPS} &= I(u, v) \times H_{FAF(u,v)} \\ &= \left\{ \left| R\left(\frac{x_2}{\lambda f}, \frac{y_2}{\lambda f}\right) \right|^2 + \left| T\left(\frac{x_2}{\lambda f}, \frac{y_2}{\lambda f}\right) \right|^2 \right. \\ &\quad + 2R\left(\frac{x_2}{\lambda f}, \frac{y_2}{\lambda f}\right) T\left(\frac{x_2}{\lambda f}, \frac{y_2}{\lambda f}\right) \\ &\quad \times \cos\left\{ 4\pi\left(\frac{a}{\lambda f}x_2 + \frac{b}{\lambda f}y_2\right) + \phi_R\left(\frac{x_2}{\lambda f}, \frac{y_2}{\lambda f}\right) \right. \\ &\quad \left. \left. - \phi_T\left(\frac{x_2}{\lambda f}, \frac{y_2}{\lambda f}\right) \right\} \right\} \left[\frac{1}{\left| R\left(\frac{x_2}{\lambda f}, \frac{y_2}{\lambda f}\right) \right|^2} \right] \end{aligned} \quad (10)$$

If the reference and the target are similar, then FJPS output can be written as:

$$\begin{aligned} \text{FJPS} &= 2 \left\{ 1 \right. \\ &\quad + \cos\left\{ 4\pi\left(\frac{a}{\lambda f}x_2 + \frac{b}{\lambda f}y_2\right) + \phi_R\left(\frac{x_2}{\lambda f}, \frac{y_2}{\lambda f}\right) \right. \\ &\quad \left. \left. - \phi_T\left(\frac{x_2}{\lambda f}, \frac{y_2}{\lambda f}\right) \right\} \right\} \end{aligned} \quad (11)$$

Resulting into dc and two correlation peaks centered at $(x_3 = 2a + 2b)$ and $(y_3 = -2a - 2b)$.



4. Results

Results of computer simulation

To investigate the performance of the proposed technique, simulation experiments are performed using MATLAB. Employing a set of fingerprint images. A 78 x 78 image was used for each reference and input. The reference and the input images were combined, along the first and third quarters respectively, on a zero padded of 1200 x 1200 pixel size forming the input joint image as shown in Figure 2 (a). The distance between the two images is necessary to avoid the overlap between the zero-order peak and the cross-correlation sidelobes peaks.

Figure 2 (a) shows the joint input in case of similar reference and target images. The joint input was Fourier transformed and the JPS was obtained. For JTC, the Fourier transformed taking to the JPS. For FJTC, the JPS was multiplied by FAF and then inverse Fourier transformed to achieve the correlation peaks. The corresponding simulation results for the conventional JTC and FJTC are shown in figure 2 (b) and figure 2 (c) respectively.

The correlation output for the classical JTC is shown in figure 2 (b) which shows that it is very difficult to detect the target since the autocorrelation outputs are almost buried in the noise floor. Since the classical JTC yields poor correlation results for images. The output consists of three terms, central order as strong dc, and two bright sharp spots diffraction orders, corresponding to two autocorrelation peaks. The correlation peaks coordinates are relevant to the

distance between two images (target and reference) and displacement the target and the reference image.

A conventional JTC usually generates many shortcomings, such as large correlation sidelobes, large correlation peak width, strong zero-order peak, and low diffraction efficiency. To inhibit noises and sidelobes the JPS is multiplied

by a FAF shown in figure 2(d) before the inverse Fourier-transform operation to obtain high cross-correlation peak correlation peak intensity, small correlation sidelobes, and narrow correlation bandwidth. For the fringe-adjusted JTC, $B(u, v)$ is taken to be unity to ensure that the FAF gain is less than or equal to unity. When compared to the classical JTC, we observe from figure 2(c) that the autocorrelation peak intensity had increased manifold and the target can be detected without any ambiguity. We see that the FJTC increases the resolution of the cross-correlation peaks providing higher peak intensities and narrower cross-correlation peak widths, and produces better discrimination between similar inputs.

Figure 3 (a) shows the joint input when both the reference and target weren't same. The corresponding simulation results for the conventional JTC and FJTC are shown in figure 3(b) and figure 3(c) respectively.

We can see from two figures, a similarity between the two images can be measured by quantifying height and sharpness of a correlation output signal. If the target image is identical to the reference image, its correlation output has a high and sharp peak. In the case of non-identical images, its correlation output doesn't have peak.

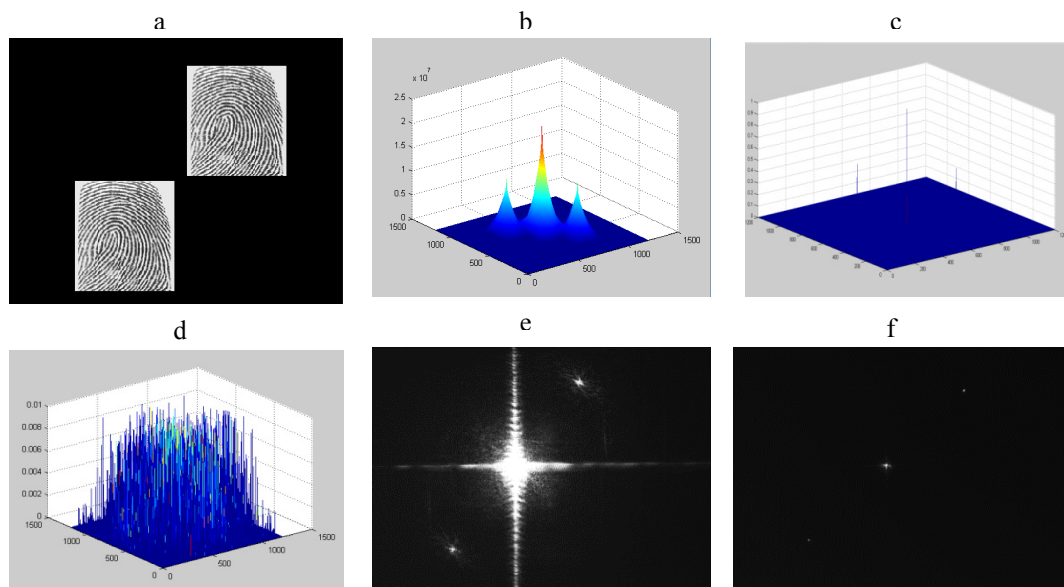


Figure 2: (a) Joint input image having reference and target same; (d) the 3D Fringe-Adjusted Filter; corresponding simulation results for the correlation output of (b) JTC, (c) FJTC; corresponding experimental results for the correlation output of (e) JTC, (f) FJTC.



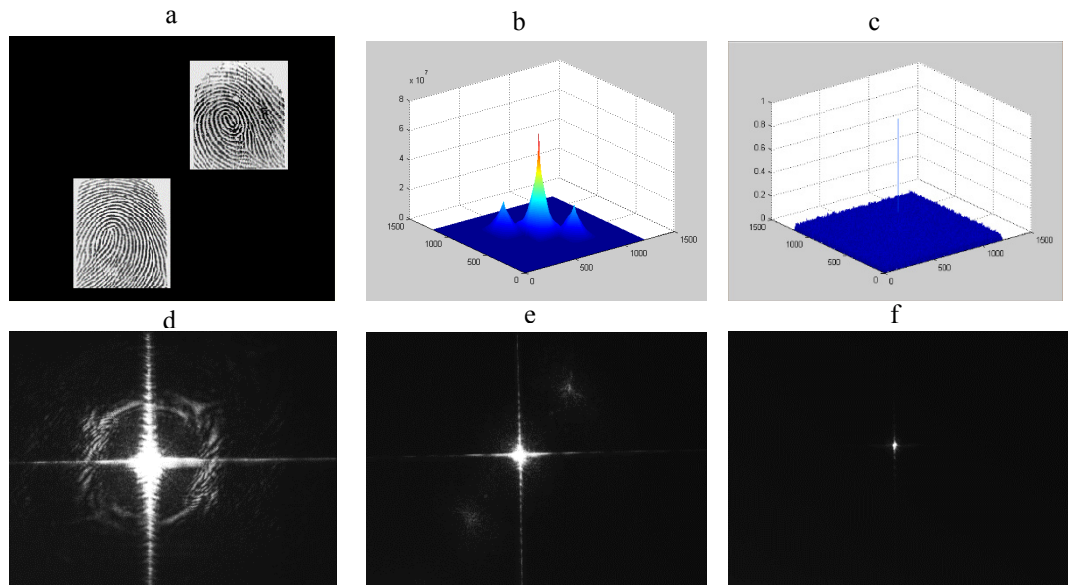


Figure3: (a) Joint input image having reference and target different; corresponding simulation results for the correlation output of (b) JTC, (c) FJTC; corresponding experimental results for the correlation output of (d) JTC, (e) FJTC.

Experimental results

The experimental setup used for implementing the JTC and FJTC is demonstrated in Figure 4. A collimated light beam from a He-Ne laser source ($\lambda=632\text{nm}$) is passed through a microscope whose function is to expand the beam; larger power (e.g., $60\times$) gives larger beam in shorter distance. This beam passes then through a CL lens with focal distance of 100mm in order to collimate the beam. The collimated laser beam is used then to illuminate Liquid Crystal-on-Silicon (LCOS) active matrix reflective mode phase only LCD with 1920×1080 resolution, 0.7 diagonal, $0.8 \mu\text{m}$ pixel pitch and 60 Hz image frame rate used to display input joint-image. A Fourier transforming lens with 150 mm focal length takes the Fourier transform. The JPS (and correlation peaks) are captured using a CCD camera (SXVR-H9C) a square-law detector with $1392 \times 1040 \times 6.45\mu\text{m}$ pixels. $6.45\mu\text{m}$ pixels in an $8.98 \times 6.71\text{mm}$ active area. From Sony, Japan Connected to a personal computer through a USB cable. For carrying out experiments, The target image and the reference image are displayed side by side on an (EASLM) placed in a front focal plane of a Fourier-transforming lens. Then the EASLM illuminating by laser, the displayed joint images are optically Fourier transformed by the lens. Then a joint power spectrum (JPS) generated at the back focal plane of the lens. The JPS is captured by the CCD and sent to computer and display in SLM to Fourier transforming optically then the correlation output is finally produced. Thus, the CCD camera captures these peaks and transfers them to the PC, and then we can act upon their values and decide the presence or the absence of the target in the input scene. To achieve the FJTC, after captured JPS and

sent to computer it is multiple by FAF and then inverse Fourier transformed.

In order to compare the simulation output with that of the experimental studies, same images used for simulation study were used for carrying out the experimental study. Figure 2 (a) where the reference and target are same was taken as the input joint image. The corresponding experimental results for JTC and FJTC are shown in Figure 2 (e) and Figure 2 (f) respectively. The JPS captured by the CCD is shown in Figure 3 (d). As shown in Figure xxx, the experimental simulations demonstrated the advantageous correlation peak produced by a FAF technique compared to that in case of conventional JTC. Achieved distinct cross-correlation peak and improved significantly the detection ability.

The joint input for reference and target are not same Figure 3 (a) was displayed onto SLM to obtain JPS. The JPS was optically inverse Fourier transformed to yield the correlation output. Figure 3 (e) shows the correlation peaks for JTC. For FJTC, the JPS digitally multiplied by FAF, known as FJPS and was then optically inverse Fourier transformed. We used the similar values for $A(u,v)$ and $B(u,v)$ as taken in the previous case. The correlation result is shown in Figure 3 (f).

5. Conclusion

The objective of this paper is to solve the intensive computational problems that requires long time in case of digital processing using Optical systems. the main advantage of optical systems is its capability to parallel perform a two dimensional correlation operations using the FT property of the lenses. In this study, We built an optical pattern recognition system



using JTC and FJTC for real-time target recognition applications. Simulation as well as experimental results proved that FJTC for similarity measurement yields substantially better correlation output than the classical JTCs. FAF is designed in such a way to avoid the problems associated with the inverse filter, as well as, to producing a very high autocorrelation peak intensity in case of similarity than in case of dissimilarity. In addition, It may attenuate the noise in the input scene provided that the factor A is selected

properly. An efficient and high-speed fingerprint verification technique is proposed. The proposed technique is observed to be very successful in recognizing a target fingerprint which is similar with reference and reject any other fingerprints.

In future, our system will be tried on a set of medical images namely retinal photographs which are used as signatures for biometric purposes.



Figure 4. Experimental set-up used

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